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ELEKTOR ELECTRONICS

THE INTERNATIONAL ELECTRONICS MAGAZINE

MARCH 1993

UK £2.25

4 Mbyte printer buffer card

Antenna test instruments

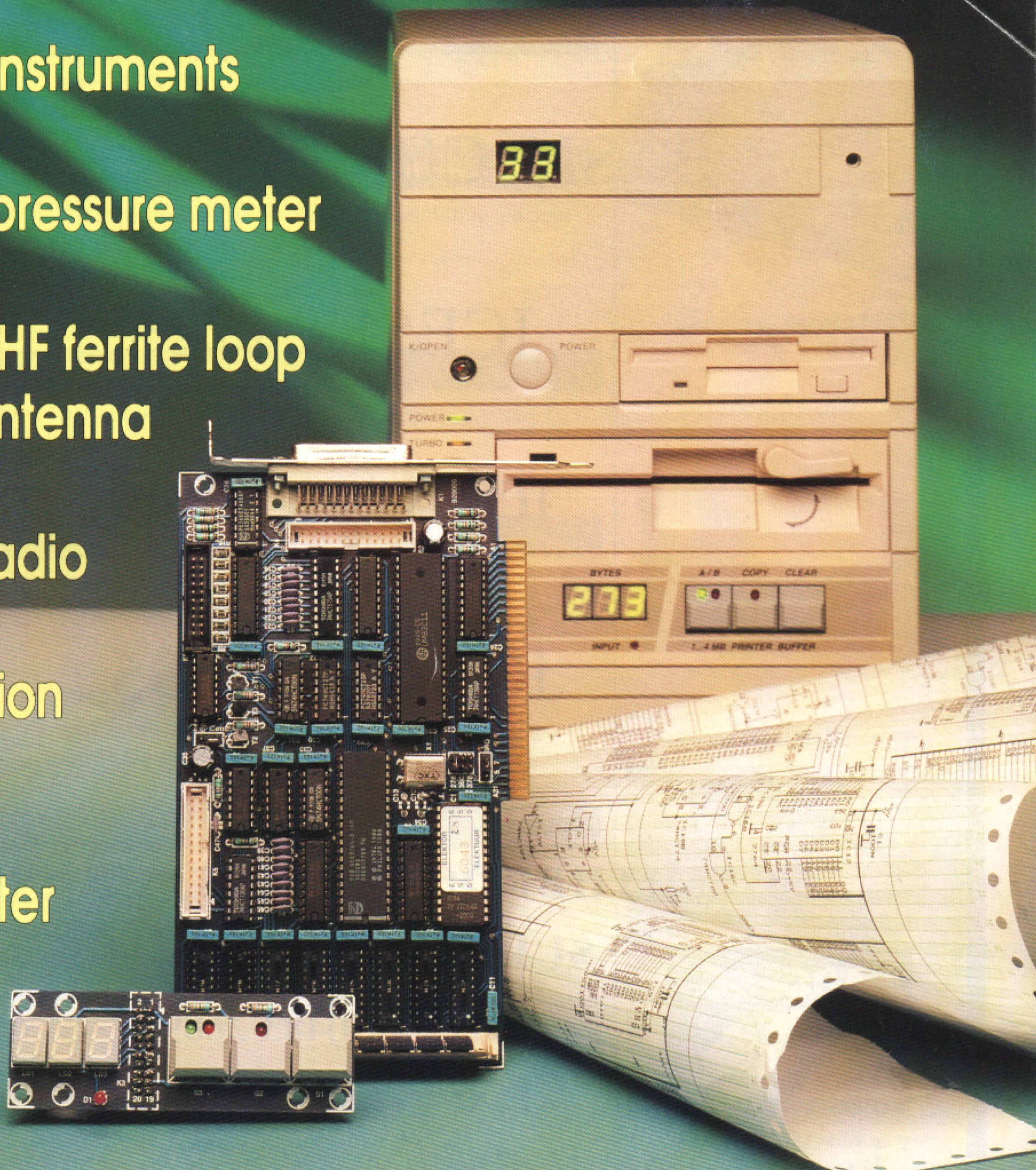
Linear sound pressure meter

Experimental HF ferrite loop transmitting antenna

World-band radio

8032 application notes

Watt-hour meter – Part 2



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ELEKTOR ELECTRONICS

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Number 209

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- Audio power meter
- A progressive and holistic design for a sound wall
- Temperature-insensitive voltage divider
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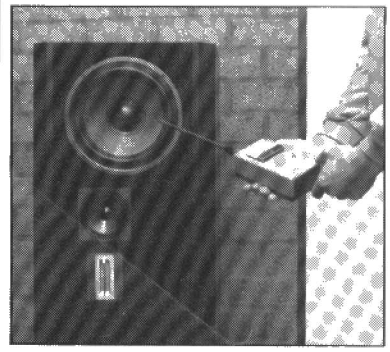
- Dynamic loudspeaker system
- DX Television (2)
- 80C32 computer application notes
- and others for your continued interest.

Front cover

Printer buffers are usually thought of as computer peripherals, that is, 'boxes' connected between the computer and the printer. Put that thought aside and you will see the advantages of the printer buffer insertion card in the article that starts in this issue on page 20. The buffer described there allows RAM ICs as well as RAM modules (SIPP or SIMM types) to be used. It offers a total buffer capacity of 1 Mbyte or 4 Mbyte, depending on the type of memory fitted. The card is suitable for use with 286, 386 or 486 driven MS-DOS machines.



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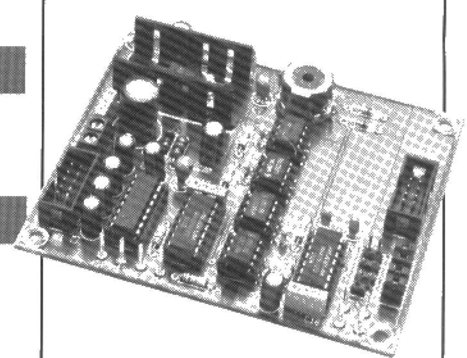

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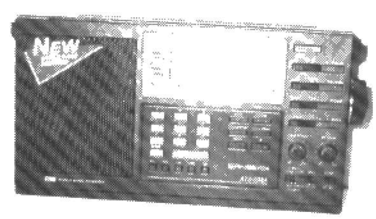

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ELECTRONICS SCENE

John G. Kemeny (1926–1992)

Dr John Kemeny, creator of the computer language BASIC, died on 26 December, 1992. He was born in Hungary.

John Kemeny arrived in the USA in 1940 as a refugee from the Second World War. He studied mathematics at Princeton and obtained his doctorate at the early age of 23. Following that, he worked as a research assistant with Albert Einstein.

At the age of 27 he was appointed professor of mathematics at Dartmouth College, where he remained until his retirement in 1990.

DIGITAL RADIO ON TEST – VIA SPACE

British Broadcasting Corporation engineers are working on a digital radio system that produces sound equal to the quality of a compact disc. The BBC is working with the British National Space Centre (BNSC) and European collaborators to speed the development and introduction of what is known as digital audio broadcasting or DAB.

As a result, DAB is being tested with a series of experimental transmissions to industry via the European Space Agency's (ESA) Olympus satellite. The BNSC is providing the BBC with access to the direct broadcasting transponder on Olympus and has arranged for the satellite beam to be swung from its Italian coverage to sweep Europe. Normally, Olympus's direct broadcasting beam radiates programmes from the Italian RAI state broadcasting service.

The BBC are using the test transmissions to research the distribution of DAB signals to terrestrial transmitter sites and to explore the potential for satellite delivery of DAB for public reception. The Olympus satellite will also be used to study aspects of picture scrambling systems and the distribution of digital television.

Although the BBC has made no decisions about DAB, reports suggest that the test system, which is being developed under the EUREKA programme of collaborative European hi-tech research, could be up and running in the second half of the decade.

DAB has the advantage that it can be received by radios in vehicles without the risk of fade-out or loss, and listeners will be able to tune directly into the satellite. The signals can also be sent from the satellite to terrestrial radio masts for re-broadcasting to receivers with a standard antenna.

MULTIMEDIA ON THE DESKTOP

The Anglo-American partnership of British Telecom (BT) and Motorola has merged video and microchip technologies to enable desktop computers and workstations to be converted into personal video communicators so that their users can hold face-to-face meetings across the world, access local and remote databases, and share information interactively.

The two organizations have joined forces to provide what is described as the first 'complete and affordable multimedia communications solution' in a new generation of advanced telecommunications products. Multimedia communications is a combination of television, facsimile, and computer links with audio and video capabilities, using a single terminal. The new system is due to come on to the market next year.

Motorola will integrate BT's standards-based video coding technology into a personal computer multimedia communications chip set capable of simultaneously processing real-time video, still images and data.

BT will incorporate this chip set into future video communications products such as slot-in computer cards. The companies expect the volume pricing for the chip set to be about £50.

FASTER PROTYPING IN PROSPECT

The design and development of new products will be speeded up considerably if a major European research and development plan to develop new computer aided rapid prototyping (CARP) technology is successful.

The three-year project is being carried out under the auspices of the EC's EUREKA high-technology programme and involves a consortium made up of a number of original equipment manufacturers (OEMs), software developers and Leeds University. It is backed financially by the Department of Trade and Industry.

CARP, described as a European research initiative, will seek to integrate computer-based design, analysis and manufacturing methods. At the core of the programme is the development of procedures for the highly resolved and accurate representation of components with the use of three-dimensional computer-aided design (CAD) models. The associate analytical and fast free form fabrication (FFFF) manufacturing technologies will be adapted to use this computer definition of component shape.

BOOST FOR SILICON SCOTLAND

Motorola will invest another £40 million in its plant in East Kilbride in Scotland to put it at the forefront of the computer chip technology industry.

New facilities being provided will enable even more circuitry to be packed on to a single chip of silicon and will reinforce Scotland's position as producer of more than half of the UK output and around 20% of all semiconductors made in Europe.

Motorola says that the move to 0.5 μm geometry by next year will transform East Kilbride into Europe's leading, fully-integrated, volume semiconductor plant. The new geometry means that a street map of the entire UK can be placed on to a piece of silicon of 0.5 cm^2.

At present, the East Kilbride plant produces 12 million chips a week for use in eveything from domestic machines to smart plastic cards, computers and automobile control systems. The new investment will enable output to be increased by 20%.

EYE MOVEMENTS MAY OPERATE COMPUTERS

Researchers in Britain believe it may soon be possible to translate a person's eye movements into computer commands. Optometrist Dr Peter Howart, who is working on the idea at Loughborough University's human sciences department, says the aim is to use the eyes to emulate a computer mouse and so allow people with limb disabilities to enter the modern world of computers.

Dr Howarth, working with ergonomist Howell Istance from the computing department at De Montford University, has produced an eye-tracking system that monitors the position of a person's eyes and sends the computer's screen cursor to wherever he or she is looking.

The binocular tracker system, based on a £50,000 tracker used by the advertising industry to assess the impact of its TV advertising, uses a pair of infrared cameras that sample eye position 60 times a second. As the eyes move left, right, up or down, the system records the eye position and sends the appropriate commands to the computers just as if it were a mouse. The screen cursor then moves around the screen as the person's eyes move.

GROWTH OF EUROPEAN CABLE TV

Europe's cable TV business is growing to the extent that it will soon change from being just another medium for entertainment to one that is capable of providing a host of services to both businesses and homes alike, according to a report from Frost & Sullivan.

The cable TV market in western Europe,

worth (in 1992) just under £410 million, is often compared with the more mature market in the USA, estimated at around £825 million. There are, however, many factors that make comparisons difficult says the report, not the least of which is the currently increasing market penetration of DTH and SMATV systems, notably in the UK and Germany.

It is clear, however, that Europe has a very long way to go before it catches up with the USA, if it ever does. Indeed, the report predicts that by 1997 western European cable TV business will have grown to about £800 million, still just short of the present US figure.

The highest growth rate is forecast to be by fibre optic cable (from about £28 million in 1992 to almost £140 million by 1997) mainly at the expense of coaxial cable for which a negative growth is predicted.

Frost & Sullivan, 4 Grosvenor Gardens, London SW1W 0DH.

FIRST WESTERN-BUILT SATELLITE TO BE LAUNCHED ON RUSSIAN PROTON ROCKET

Inmarsat will be the first western-based organization to use a Russian Federation launcher to place one of its satellites into orbit. The launch will be of the fourth of the new Inmarsat-3 satellites and will take place in late 1995 on a Proton rocket from the Baikonur Cosmodrome in Kazakhstan in the CIS.

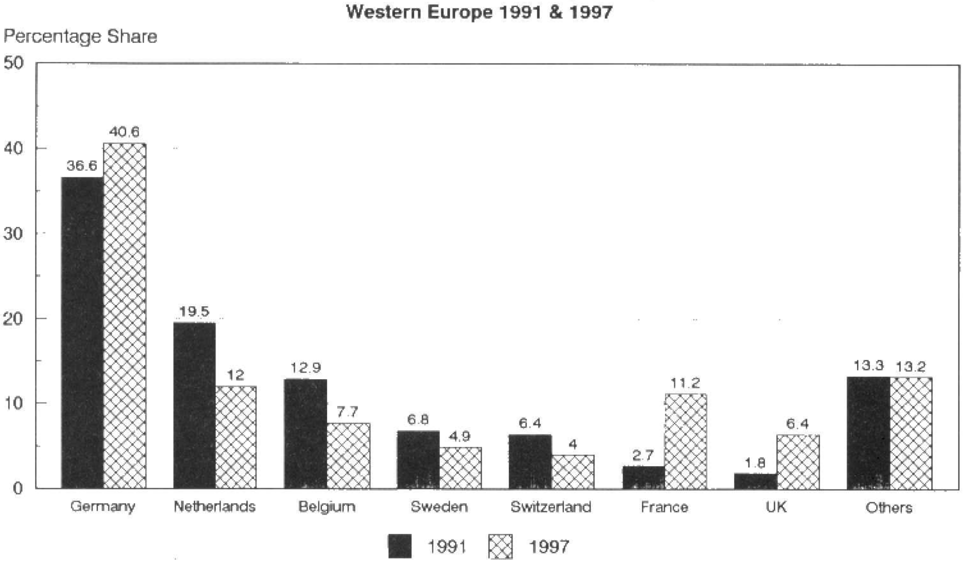
The launch in 1994 of the first two of the new satellites was awarded to General Dynamics, while the third, scheduled for mid-1995 will be on an Ariana rocket.

The mix of launchers minimizes the risk for Inmarsat in what is demonstrably a high-risk business.

The new satellites will use the latest spot-beam technology to provide voice and data communications services worldwide to mobile terminals as small as pocket-size, as well as on ships, aircraft and vehicles. They will have about eight times the power of Inmarsat's current generation, making it possible for users to communicate via smaller and cheaper mobile terminals and possibly resulting in an eventual decrease in user charges.

Inmarsat, 40 Melton Street, London NW1 2EQ.

Share of CATV Subscribers by Country
Western Europe 1991 & 1997



Source: Frost & Sullivan
Report E1645

MAPLIN NICAM STEREO RECEIVER

With NICAM transmissions now a familiar part of British TV transmissions, Maplin are making it possible for your standard TV to reproduce near-CD-quality stereo sound. The new receiver may be connected to the hi-fi system in the same manner as a radio tuner or CD player. It converts the digital sound track transmitted alongside the normal mono FM sound and picture from TV stations in the UK.

Full details can be found on p. 255 of Maplin's 1993 catalogue, which is available from W H Smith or your local Maplin shop at £2.95 or by mail order from Maplin at £3.45.

Maplin Electronics, P.O. Box 3, Rayleigh, Essex SS6 8LR. Phone 0702 554 161.

EUROPE'S FIRST NEWS CHANNEL ON EUTELSAT

Euronews, the first European news and information satellite TV channel was launched on EUTELSAT II-F on 1 January.

The new channel is broadcast from 0600 to 0200 CET. Accompanying sound is available simultaneously in English, French, German , Italian and Spanish.

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An Arabic soundtrack is planned for the middle of next year.

The new channel can be received on transponder 37 of EUTELSAT II-F1 at 13° E, vertical polarization. Downlink frequency is 11.575 GHz with sound in German on 6.65 MHz, English on 7.02 MHz, Spanish on 7.20 MHz, French on 7.38 MHz, and Italian on 7.56 MHz.

Euronews estimates that it is reaching 35 million homes connected to cable networks or with domestic satellite reception systems. Direct reception via EUTELSAT II widebeam channels is possible with 80 cm antennas in all of central and western Europe, and with slightly larger antennas in northern Scandinavia, eastern Europe and North Africa.

EUTELSAT, Tour Maine-Montparnasse 33, avenue du Maine, 75755 Paris, Cedex 15, France.

EUTELSAT TV LINE-UP

Channel	Tr.	Freq.	Audio / Language	Beam
EUTELSAT II-F1 - 13°E				
ARD	33H	11.596	7.02, 7.20 - German	Superbeam
Avrasya	22H	11.181	6.65 - Turkish	Superbeam
Der Kabelkanal	21H	11.055	(D2 Mac) - German	Superbeam
Deutsche Welle TV	27V	11.163	6.65 - German, English, Spanish	Widebeam
Eurosport	20H	10.972	6.65, 7.20 - German, 7.02 - English 7.38 - Dutch, 7.56 - French	Superbeam
Filmnet 24*	34H	11.638	6.50 - English, Dutch	Superbeam
Filmnet 24*	34H	11.678	6.50 - English, Dutch	Superbeam
MBC	32H	11.554	6.60, 7.02, 7.20 - Arabic	Superbeam
Red Hot Dutch**	22H	11.181	7.02 - English	Superbeam
Super Channel	25V	10.987	6.65 - English, 7.02 - Dutch 7.20 - German	Widebeam
TV5 Europe	26V	11.080	6.60 - French	Widebeam
Worldnet	27V	11.163	6.65 - English	Widebeam
EUTELSAT II-F2 - 10°E				
Interstar	38V	11.617	6.60 - Turkish	Widebeam
Rai 1***	26V	10.972	6.60 - Italian	Widebeam
Rai 2***	25V	11.095	6.60 - Italian	Widebeam
Show TV	37V	11.575	6.65 - Turkish	Widebeam
Teleon	33H	11.596	6.65 - Turkish	Widebeam
TVE International	22H	11.149	6.60 - Spanish	Widebeam
EUTELSAT II-F3 - 16°E				
Antena Tres°	25V	10.972	6.60 - Spanish	Widebeam
Canal Plus Espana°	27V	11.178	6.60 - Spanish	Widebeam
Croatian TV	20H	10.987	6.65 - Serbo-Croat	Superbeam
Duna 7	33H	11.596	6.50 - Hungarian	Superbeam
ESC (Egyptian Satellite Channel)	22H	11.163	6.60 - Arabic	Superbeam
HBB	38V	11.617	6.65, 7.02/7.20 - Turkish	Superbeam
Polsat	34H	11.678	6.60 - Polish	Superbeam
RTP International	37V	11.575	6.60 - Portuguese	Widebeam
Tele Cinco°	26V	11.095	6.60 - Spanish	Widebeam
TV7 Tunisia	39V	11.658	6.60 - Arabic	Widebeam
TV Plus°°	34H	11.678	D2 Mac - Dutch	Superbeam
EUTELSAT II-F4 - 7°E				
ET1	22H	11.178	6.60 - Greek	Widebeam
Kanal 6	27V	11.163	6.60 - Turkish	Widebeam
RTS Sat	34H	11.638	6.65 - Serbo-Croat	Widebeam
RIK	22H	11.144	6.60 - Greek (Cypriot)	Widebeam

For up to the minute information on EUTELAT 24 hours a day, call OLE (On - line EUTELSAT) on (33) 1 43 21 23 38 and receive news by phone or fax.

* Matsushita Satbox encryption; ** SAVE; *** Discret 12;
° Nagravision; °° Eurocrypt VOC/12/92

Courtesey European Telecommunications Satellite Organization.

KITS FROM CANAL BRIDGE AUDIO

Canal Bridge Audio have available a full range of kits and professionally finished products for electronics enthusiasts. Customers will have the choice between top-quality, reasonably priced kits, modules and professionally finished products. Among the many interesting items is a recording briefcase which, with a built-in sensitive microphone, is unparalleled in the making of high-quality recordings of business meetings, conferences, and so on.

Moreover, an advice service is provided by experienced staff ready to help with all manner of queries and giving full technical back-up.

For a full catalogue, write to **Canal Bridge Audio, 172 Caledonian Road, London N1 0SG. Phone 071 837 4423.**

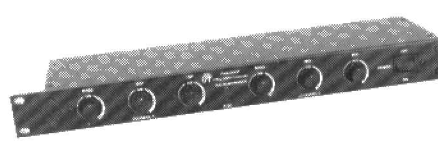
MMIC RF SWITCHES FROM CIRKIT

Cirkit now stock Alpha Industries' series of low-cost GaAs FET MMIC switches and attenuators. These surface mount devices, in plastic SOIC packages, are particularly suited for cellular phone applications.

The AS002M2-12 is an SPDT RF switch operating at up to 2.5 GHz. The switches can be arranged in series for multiple throws and for higher isolation applications. Other types in the range include double-pole switches, four-way switches, attanuators and control FETs.

Cirkit Distribution, Park Lane, Broxbourne, Herts EN10 7NQ. Telephone (0992) 444111.

B.K. GO ACTIVE



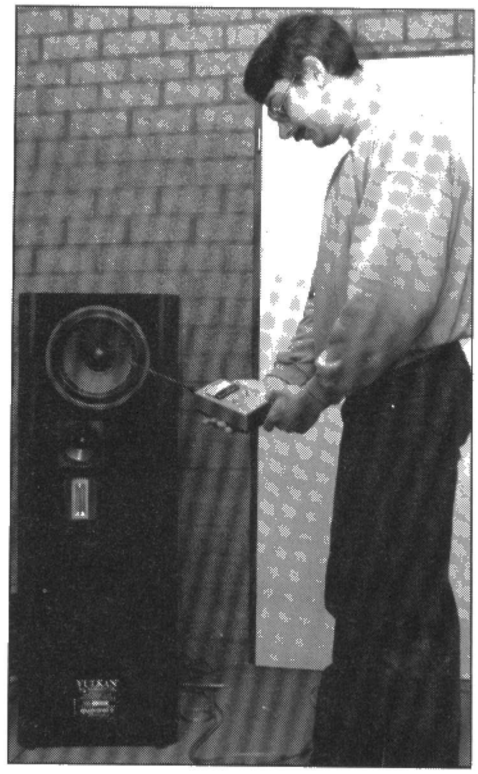
B.K. have launched a high-quality programmable, 3-way stereo cross-over unit, the XO3. Housed in an industry-standard 19 in. 1U high rack case, the new unit has a removable front fascia panel, located behind which are the DIP switches for programming the cross-over points. Levels for bass, mid and top are fully adjustable, with phase invert switches on the bass channels. The XO3 achieves a 24 dB per octave cross-over slope.

The unit is available at £116.33, incl. VAT, plus £7.00 delivery from **B.K. Electronics, Units 1&5 Comet Way, Southend on Sea SS2 6TR. Telephone (0702) 527572**

LINEAR SOUND PRESSURE METER

Design by T. Giesberts

It is often next to impossible to determine the acoustic behaviour of a home-built loudspeaker enclosure other than by ear. The meter described here enables the frequency characteristic of a loudspeaker system to be ascertained.



The sound pressure meter is based on a measurement (or reference) microphone that is usable up to 20 kHz and a sound source. The sound source may be a wobbulator or, if that is not available, a measurement CD (compact disc). A measurement CD provides up to 30 noise frequencies at intervals of a third of an octave.

Wobbulating, that is, the simultaneous generating of several different test frequencies in a narrow band, serves to nullify the effects of the room in which the loudspeaker system is used. Usually, the output of the wobbulator shifts between two frequencies that are separated by, say, a third of an octave.

A measurement CD produces pink noise (noise of which the amplitude is inversely proportional to the frequency), which has been filtered in a way that leaves only certain frequencies.

The microphone

As will be realized, the quality of the sound pressure meter is highly dependent on the quality of the measurement microphone. It is not necessary that the frequency characteristic of the microphone is absolutely straight from 20 Hz to 20 kHz. However, its characteristic must be known precisely so that it can be corrected electronically. For the present design, a number of microphones were tested, of which two were found to be the most suitable. One is the Sennheiser Type KE4-211-2, which was found to be particularly good at the higher frequencies. As far as can be ascertained, this retails at £25–£30. Somewhat cheaper is Monacor's Type MCE-2000 (about £8–£10).

These microphones were compared with a Bruel & Kjaer Type 4155 reference microphone: the deviations of their characteristic from that of the 4155 are shown in Fig. 1. The characteristics show the difference in signals between the reference microphone and the tested microphone.

The slight upward movement of the Sennheiser characteristic arises from the smaller diameter of this microphone, which makes it more sensitive omni-directionally so that it picks up more reflections.

Although the characteristic of the Monacor type shows more undulations, it should be borne in mind that for measurements in a normal domestic room (where spurious reflections are legion) this does not matter all that much.

The slight fall-off in the Sennheiser characteristic at low frequencies is compensated in the design by a passive *RC*-network (R_2C_2), resulting in –3 dB points at 20 Hz and –0.5 dB at 50 Hz.

The production spread of frequency tolerances, tested on several of each type of microphone, was found to be within 1 dB over the audio range. Larger variations were found in the output levels, but that is of importance only for the measurement of absolute sound pressure.

Circuit description

Although the circuit in Fig. 3 is straightforward, all possible measures have been

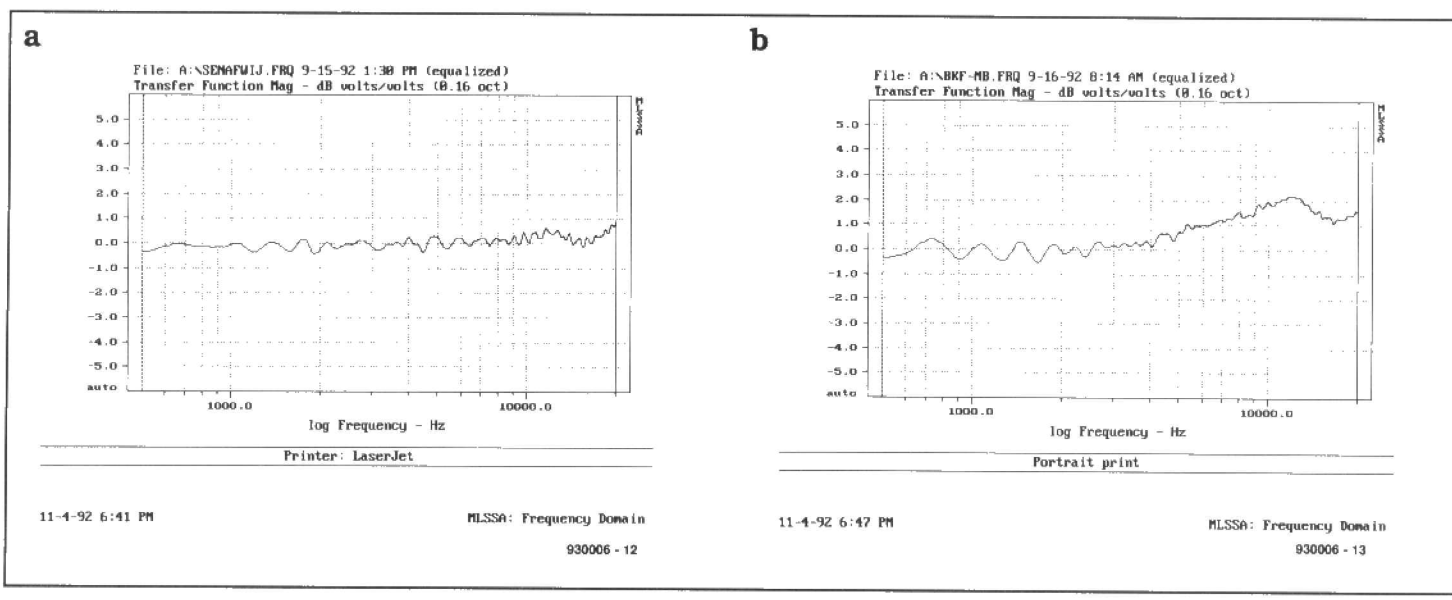


Fig. 1. Deviation of the characteristic of a Sennheiser KE4-211-2 microphone (a) and a Monacor MCE-2000 microphone (b) from a Bruel & Kjaer Type 4155 reference microphone.

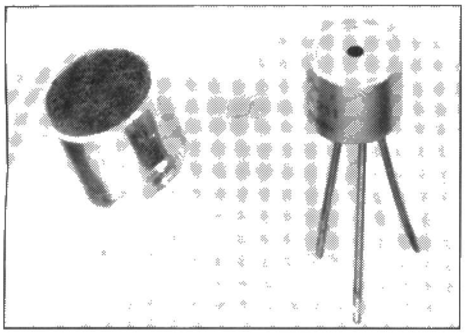


Fig. 2. Two suitable microphones: the Sennheiser Type KE4-211-2 and the Monacor Type MCE-2000.

taken in the design to obtain good linearity and accuracy.

The FET in the microphone is used as an amplifier: R_2 and R_3 serve as drain resistors. Network R_2C_2 provides compensation at low frequencies if the Sennheiser microphone is used. Network R_1C_1 decouples the supply line to the microphone. The input signal, amplified by ×3–×5, depending on the type of microphone used, is applied to opamp IC_1.

The amplification of IC_1 and IC_2 is determined by the position of switch S_1. In position '120 dB', R_6 and R_7 are short-circuited. IC_1 then functions as a voltage follower, and IC_2 amplifies by ×1, because the value of R_{10} plus R_8 in parallel with R_9 is equal to that of R_{11}. This means that the combined amplification of IC_1 and IC_2 is ×1.

In position '100 dB', the amplification of IC_1 is determined by the ratio of R_5+1 to R_6 in parallel with R_7, which, with values as shown, is 10. Since the amplifcation of IC_2 remains ×1, the combined amplification is ×10.

In position '80 dB', the amplification of IC_1 remains ×10. Since R_8 and R_9 are now short-circuited, IC_2 also amplifies ×10. The total amplification becomes, therefore, ×100.

Part of the amplified signal may be taken from K_1 for inspecting it on an oscilloscope or feeding it to an amplifier. Also, an active (A) filter may be added between K_1 and D. This is not used in the present design, in which the output of IC_2 is taken directly to true-RMS-converter IC_3. This IC converts the alternating input to a direct current that is directly proportional to the effective level of the input. Also, it provides an output that delivers 3 mV per dB voltage variation at the the input. This makes it possible for a standard moving coil (MC) meter to be given a linear dB scale.

The converter is followed by a buffer-amplifier, IC_{4b}, that drives the MC meter. The amplification provided by IC_{4b} makes it possible for meters with a sensitivity of 30–100 μA to be used.

Negative temperature coefficient (NTC) resistor R_{15} in the feedback loop of IC_{4b} provides temperature compensation for IC_3. It is coupled to the housing of the converter with the aid of a brass 'ring'.

The full-scale current is matched to the MC meter by P_2. The value of this preset has been chosen to ensure that the meter has a range of 30 dB f.s.d., and that the ranges overlap one another by 10 dB.

The supply for the circuit is stabilized by D_3, a micropower integrated voltage reference diode, which provides a reference voltage of 2.5 V. Opamp IC_{4a} converts this into a pseudo-symmetrical voltage.

The 'zener' voltage of D_3 is amplified ×0.825 (ratio R_{20}:R_{19}), so that the DC output of IC_{4a} is 2.1 V relative to its +ve input (pin 3). Adding to this the voltage across D_3, that is, betweenthe +ve supply line and pin 3 of IC_{4a}, a stabilized voltage of 4.6 V is obtained between the +ve

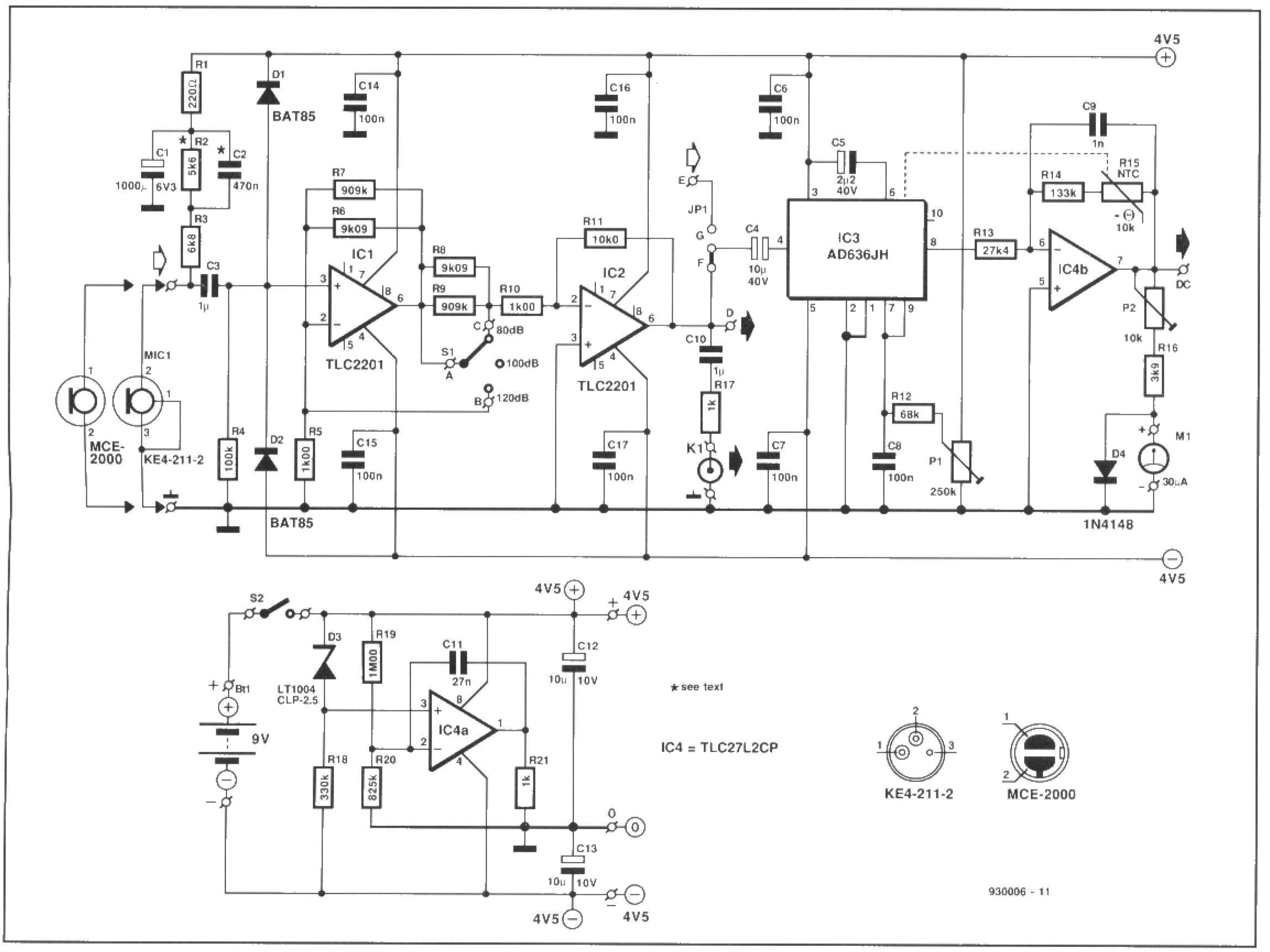


Fig. 3. Circuit diagram of the linear sound pressure meter.

supply line and the output of IC_{4a}.

The negative side of this supply may vary with the battery voltage, but this does not affect the circuit.

The stabilized supply powers the microphone, so that the microphone signal does not change with varying battery voltages.

The stabilized supply is also required for a stable I_{ref} through R_{12} to ensure that the meter scale does not vary when the battery voltage drops.

The supply voltages have been chosen deliberately at different levels to have available as high a voltage for the microphone as possible. The TLC2201s, as well as the AD636, operate satisfactorily with supply voltages down to ±2.5 V. This means that the circuit operates normally with a battery voltage of 7 V; below that, the battery must be renewed.

Construction

The circuit is best constructed on the printed-circuit board shown in Fig. 4. For the present design, a wire bridge must be laid between F and C_4. If a Sennheiser microphone is used, the values of R_2, R_3 and C_3 are as shown. If a Monacor microphone is used, replace R_2 by a wire bridge and omit C_2.

Good thermal coupling between R_{15} and IC_3 is obtained by a 3 mm copper ring sawn off a length of 12.5 mm water pipe. Squash this ring slightly and, with IC_3 and R_{15} held tightly together, force it over the assembly as shown in Fig. 5.

The microphone capsule must be mounted in a thin metal tube not shorter than 10 cm (4 in)—see photo on page 8. This is to prevent errors at high frequencies caused by the enclosure that houses the PCB, battery and meter. For the prototype a metal curtain tube was used. The capsule must be fastened at one end of the tube with superglue or similar. Before that is done, a screened cable must be soldered to it for connection to the board. Make sure that the connections are well insulated to prevent short circuits. The capsule-tube assembly may be fixed to the enclosure or be used by itself. In the latter case, it may be advisable to fit a handle of some sorts and a (XLR) connector at the end away from the capsule to make the assembly look like a real microphone housing.

The current drawn by the sound pressure meter varies between 3 mA and 8 mA, depending on the extent of the meter deflection. Therefore, if the instrument is used only occasionally, the battery will last a long time.

The instrument is calibrated by connecting an alternating signal at a frequency of about 1 kHz to the input. This signal may be obtained from a function generator or from a measurement CD. Many such CDs contain test tones that may be taken from the tape output of the relevant amplifier to the sound pressure meter. Do not yet connect the microphone to the board. Set S_1 to 120 dB.

If a Sennheiser microphone is to be used, set the input voltage to 16.5 V_{RMS} measured with a digital voltmeter across R_4. Turn P_1 so that the pointer of the MC meter is at exactly 0. Increase the input to 522 mV_{RMS} and adjust P_2 so that the meter reads 120 dB (full scale).

If a Monacor microphone is used, the procedure is the same, but the input voltages must be 8 mV_{RMS} and 253 mV_{RMS}

PARTS LIST

Resistors:
R1 = 220 Ω
R2 = 5.6 kΩ
R3 = 6.8 kΩ
R4 = 100 kΩ
R5, R10 = 1.0 kΩ, 1%
R6, R8 = 9.09 kΩ, 1%
R7, R9 = 909 kΩ, 1%
R11 = 10.0 kΩ, 1%
R12 = 820 kΩ
R13 = 27.4 kΩ, 1%
R14 = 133 kΩ, 1%
R15 = NTC, 10 kΩ
R16 = 3.9 kΩ
R17, R21 = 1 kΩ
R18 = 330 kΩ
R19 = 1.0 MΩ, 1%
R20 = 825 kΩ, 1%
P1 = 250 kΩ preset
P2 = 10 kΩ preset

Capacitors:
C1 = 1000 µF, 63 V, radial
C2 = 470 nF, see text
C3, C10 = 1 µF
C4 = 10 µF, 40 V, radial, bipolar
C5 = 2.2 µF, 40 V, radial, see text
C6–C8, C14–C17 = 100 nF
C9 = 1 nF
C11 = 27 nF
C12, C13 = 10 µF, 10 V, radial

Semiconductors:
D1, D2 = BAT85
D3 = LT1004CLP-2.5 or LT1004CZ-2.5
D4 = 1N4148
IC1, IC2 = TLC2201
IC3 = AD636JH
IC4 = TLC27L2CP

Miscellaneous:
JP1 = 3-way terminal block with jumer
K1 = audio socket
S1 = change-over toggle switch with centre position
S2 = switch with make contact
M1 = moving coil meter, 30 µA
BT1 = 9 V battery (PP3) with connector
MIC1 = microphone capsule, Sennheiser KE4-211-2 or Monacor MCE-2000
Enclosure, e.g. OKW 9030 087 or 9030 065
PCB Type 930006 (see p. 70)
Measurement CD, e.g. 'The Test', AXCD92001, from Stax, or 'Compact Test', PV784031 from Pierre Verany; available from most good audio & hi-fi component retailers

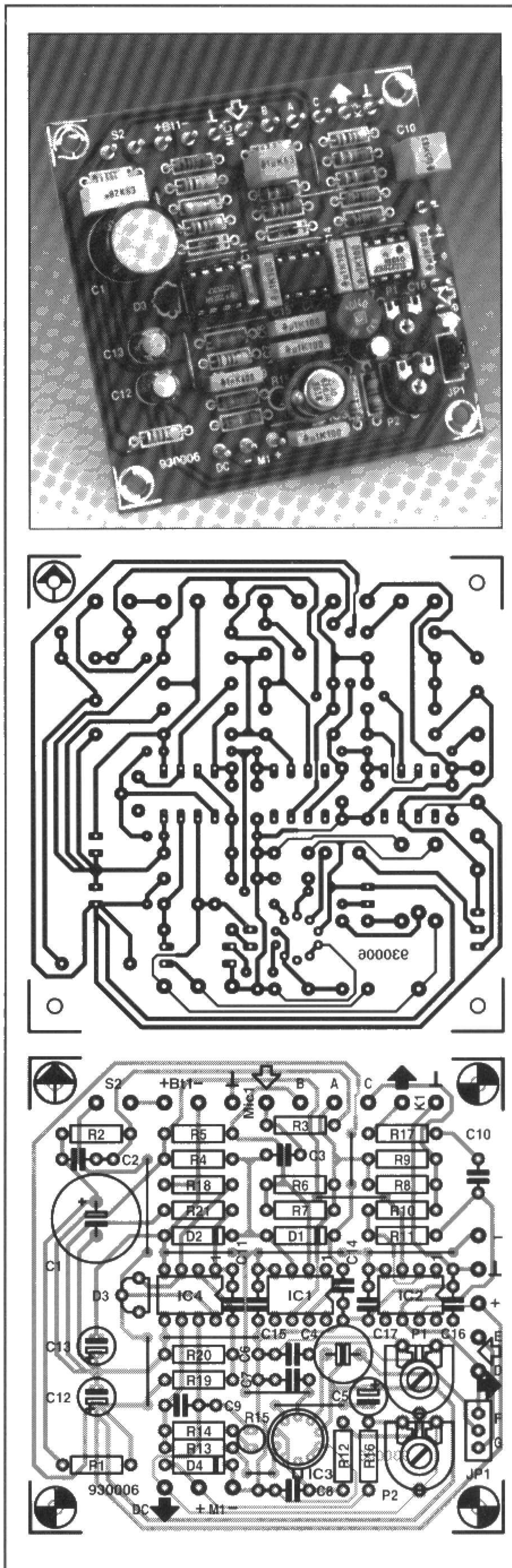


Fig. 4. Printed-circuit board for the linear sound pressure meter.

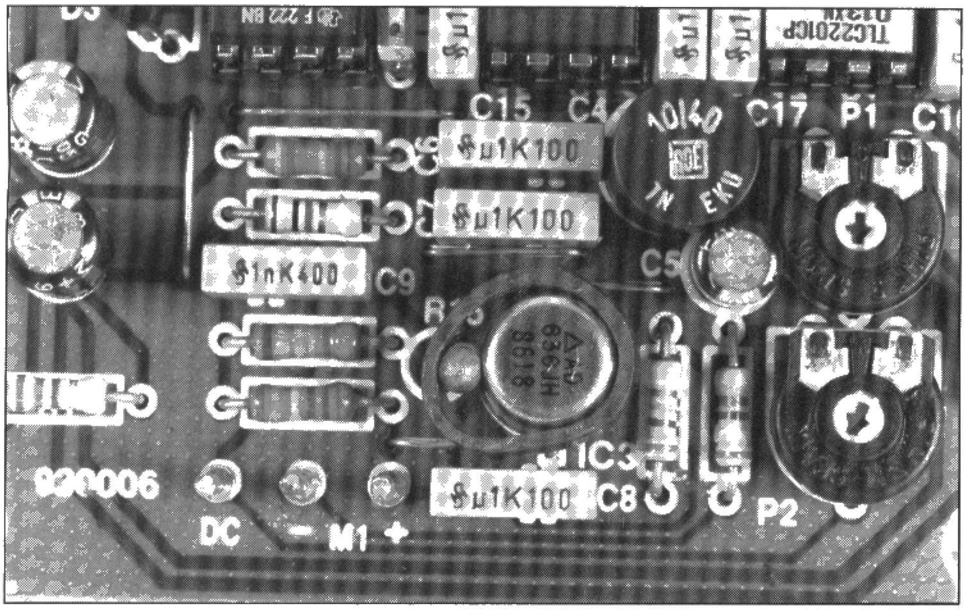


Fig. 5. The NTC resistor and converter IC must be coupled thermally with the aid of a short length of 12.5 mm (0.5 in) copper water pipe.

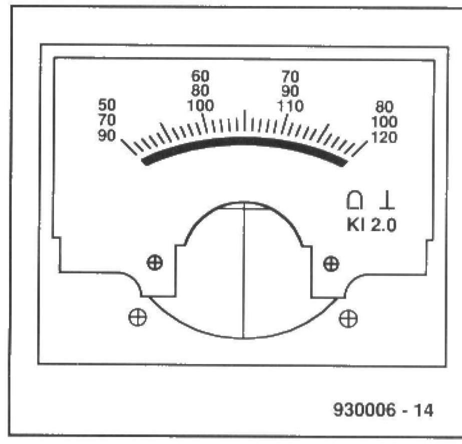


Fig. 6. The new scale for the moving coil meter.

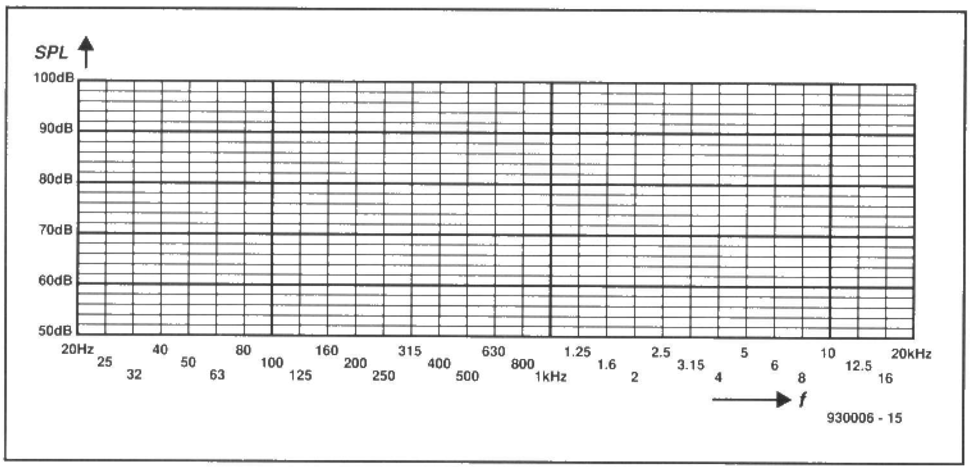


Fig. 7. Suggested layout for drawing frequency characteristics.

respectively.

When the sound pressure meter has been calibrated, the meter readings will be accurate within about 2 dB. Note that the pointer may give a negative reading if the sound pressure is below the bottom limit of the chosen scale range. That is correct and does no harm.

Measuring sound pressure

To obtain reliable characteristics, the loudspeaker enclosures must be free-standing at a height about half-way between floor and ceiling. Place the sound pressure meter at a distance of about 1 m (3 ft) from the loudspeaker at the height of the medium-range speaker or tweeter. Set the sound pressure to give a reading of 85–90 dB. Then, note down in a copy of Fig. 7 the reading for each of the 30 frequencies—see first paragraph. In many cases, measurements below about 200 Hz are prone to errors owing to the effects of the room. The only remedy for this is to take close-up measurements by placing the microphone right in front of the woofer. Unfortunately, it is difficult to correlate the two characteristics so obtained.

The characteristics at your favourite listening position can be obtained by placing the sound pressure meter there.

It is also interesting to make separate measurements on the two loudspeaker enclosures. Quite large differences are often detected, pointing to an asymmetrical positioning of the boxes.

If the MC meter varies too much during the measurements, the pointer can be made to move more slowly by giving C_5 a value of 10 µF instead of 4.7 µF. ■

AN EXPERIMENTAL HF FERRITE LOOP TRANSMITTING ANTENNA

By Richard Q. Marris, G2BZQ

Introduction and history

It had always seemed logical to the writer that when ferrite rods first appeared in broadcast receivers, the technique ought to be usable for transmission antennas.

Much of the 1970s was spent living, working and operating (G2BZQ/WO) in the USA. As an executive in the electronics industry, the writer had the advantage of a continual round of contacts with, and visits to, electronics and communications organizations in most areas of the USA. Moreover, for an amateur TX-er there was a thriving industry and unlimited supplies of equipment and components at prices apparently far more attractive than in the UK. There were several manufacturers and suppliers of ferrite rod in various materials and sizes. On the rare occasions when leisure time was available, personal experiments were carried out with various types of rod to produce external ferrite rod antennas for reception of the HF bands. After a while this led to initial experiments with ferrite rods as the basis of transmitting antennas. The 3.5 MHz band was selected and has been used ever since to compare results from experiment to experiment. It was quickly discovered that one had innocently stepped into a minefield. The use of ferrite rods for transmitting antennas was found to be a 'grey area' with an almost complete lack of information and even some mis-information. Even now, nearly 20 years later, little seems to have changed.

As personal experiments proceeded over the years, without any textbook information, it was possible to select the best available ferrite materials and produce some basic ground rules for circuitry and construction of such antennas, and this is still proceeding. Furthermore, in the early 1970s, during professional activities in the USA, rumours were heards that research and development of ferrite transmitting loops was taking place, ostensibly under Defense sponsorship, but attempts to discover details failed.

However, years later, back in the UK, it was decided to publish the results of a couple of personal experiments[1,2] with a view to identifying other amateurs who are experimenting, and also encourage others to do so. A lot of interest was raised in various countries, especially with the second design. Two amateurs were heard of who had tried actual designs and these apparently used transistor radio type (MW/LW) rods—in other words, unsuitable core materials.

The 16th Edition of the A.R.R.L. Antenna Book[3] states that ferrite loaded transmitting loops are still under development! That was in 1991!

In early 1992, Roberto Criaghero (I1ARZ) obtained, and kindly sent on, a copy of an IEEE paper by R. DeVore and P. Bohley, written in 1976[4]. These researchers, working at Ohio State University under a Defense Contract, had evaluated, over a period in the 1970s , the possible use of ferrite rods in transmitting loop antennas. This paper appears to confirm the rumours mentioned earlier which the writer had heard in the 1970s. It was most interesting, and encouraging, to note that, though their approach and configurations were quite different from the writer's, many of their conclusions regarding suitable materials and problems were the same or similar. This was rather gratifying, as the writer had plodded along, quite independently, over the years, on a trial an error bassis, without laboratory facilities, manpower, much encouragement or financial backing.

Over the years, there have been other unmentioned conflicting instances, and the reader will see that it is all a somewhat confusing subject, and a minefield for the innocent.

Summary of ferrite rod materials

Ferrite materials, suitable for ferrite rods, fall into two main categories: manganese-zinc and nickel-zinc mixtures. Manganese-zinc rods are normally used for frequencies of 1–1000 kHz and nickel-zinc rods from 200 kHz into the VHF band. There are many diferent mixtures of nickel-zinc material; Amidon Associates have produced an interesting booklet[5] covering the subject. Alas, with no reference to transmitting!

The experimenter is limited in his choice to those materials available in rod form in small quantities. Some manufacturers will produce special rods in other mixes, but only in production quantities. Over the years, tests have settled on the Type 61 nickel-zinc, which is quoted as being usable in resonant circuits from 0.2 MHz to 15 MHz, although the writer has used it successfully in the VHF band for reception purposes. Coincidently, De Vore and Bohley also set-

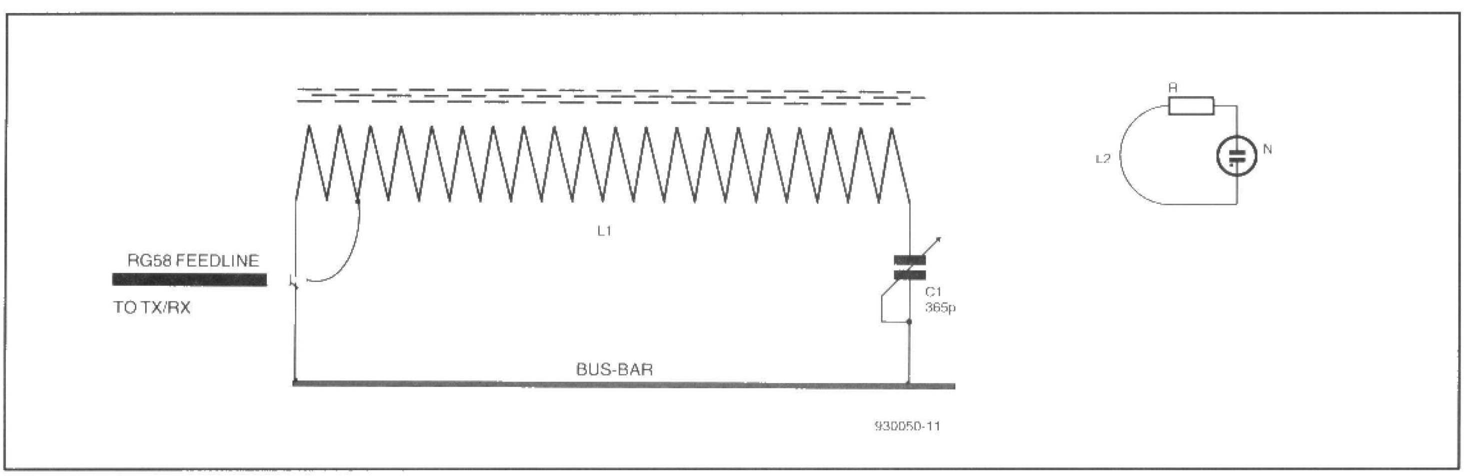


Fig. 1. Circuit diagram of the ferrite loop transmitting antenna.

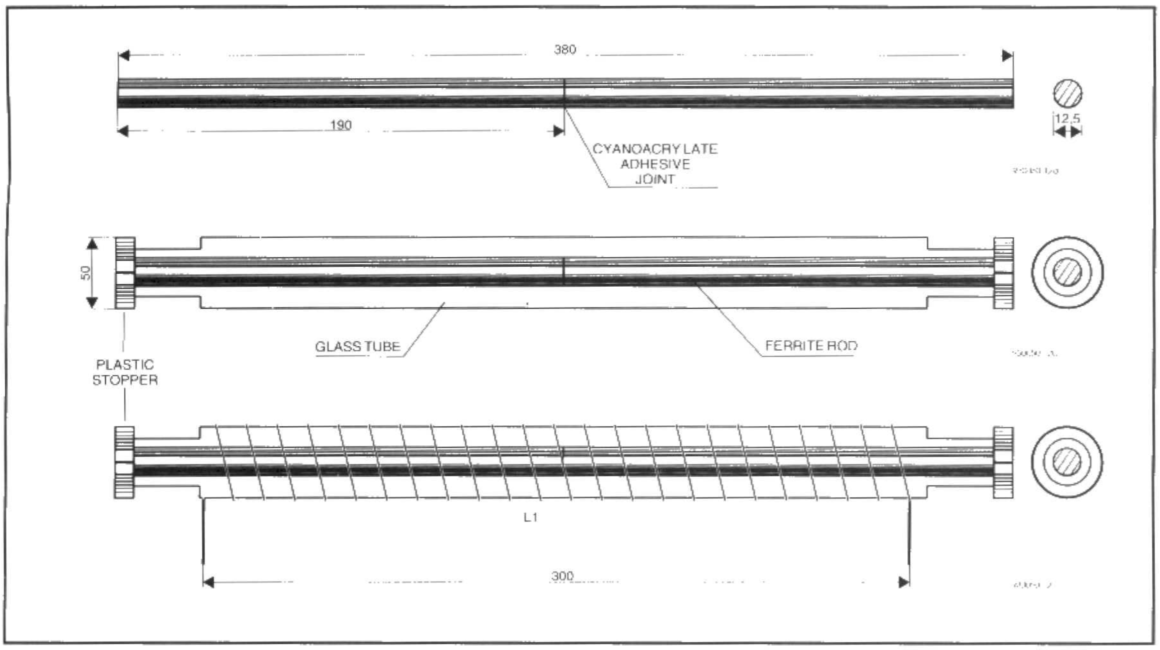


Fig. 2. Coil assembly.

tled on the Type 61 material for their tests. Type 61 nickel-zinc rods are available direct from Amidon. They are quite expensive when compared with the usual MW/LW receiver rods: their (airmail) delivery is around two weeks.

Any attempts to use ferrite rods from/for transistor radios will lead to failure.

Guide lines for using HF ferrite rods in transmitter antennas

Assuming the selected rod is usable at the required frequency, we are faced with the two interconnected main problems of 'how to couple the resulting antenna to the transmitter' and 'core saturation'. It is easy to wind a few turns of wire around a ferrite rod and resonate it to a frequency with a capacitor; add a few coupling turns and use it as a receiving antenna. This is well documented. Connect the same antenna to a transmitter, and it is unlikely to load or, if partial loading does occur, core saturation takes place.

Core saturation can easily be identified by a rapid fall-off of radiated signal; resonant-frequency drift; general instability; and in extreme cases by core and wire heating.

For transmitting antennes, the best winding configuration appears to be:

1. wire turns evenly spaced to cover the entire length of the rod;
2. sufficient air space between the wire turns and core;
3. a heavier wire gauge than used with a receiving rod antenna;
4. treat it as a low power device.

There are, however, many variables.

Over the years, many methods of matching/coupling the antenna to the transmitter have been tried. To date, the rather novel design described in this article has proved to be the best by far and has presented few loading problems.

Circuit description

The circuit in Fig. 1 shows that the antenna consists of a large diameter ferrite cored inductor, L_1, resonated at one end by variable capacitor C_1 and fed at the other end by a length of 50-Ω coaxial feedline. It is tapped at 2 turns from the feedline end. A $^1/_8\times{}^1/_8$ in (3×3 mm) brass bus-bar strip is used for the common

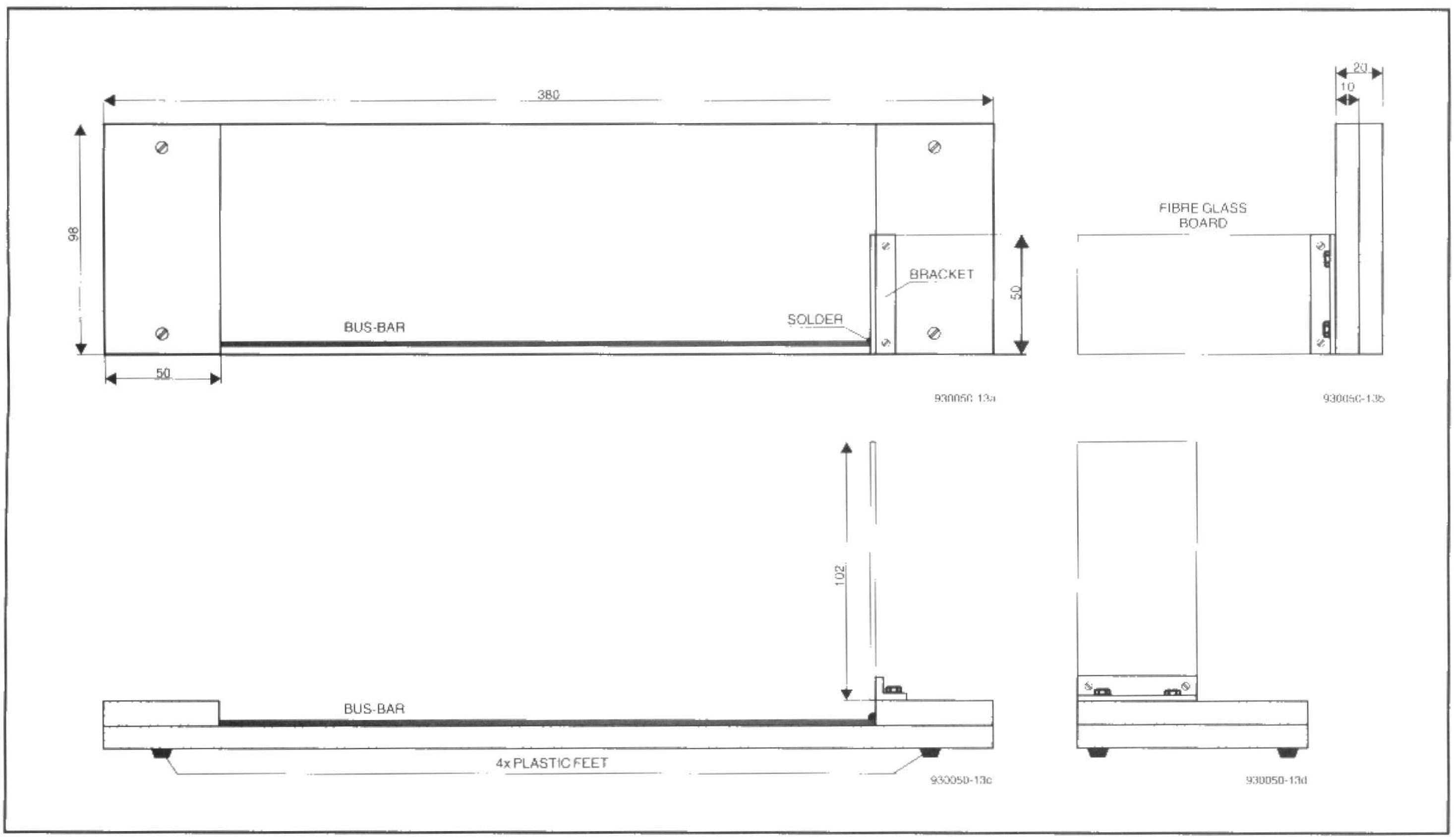


Fig. 3. Mounting base assembly.

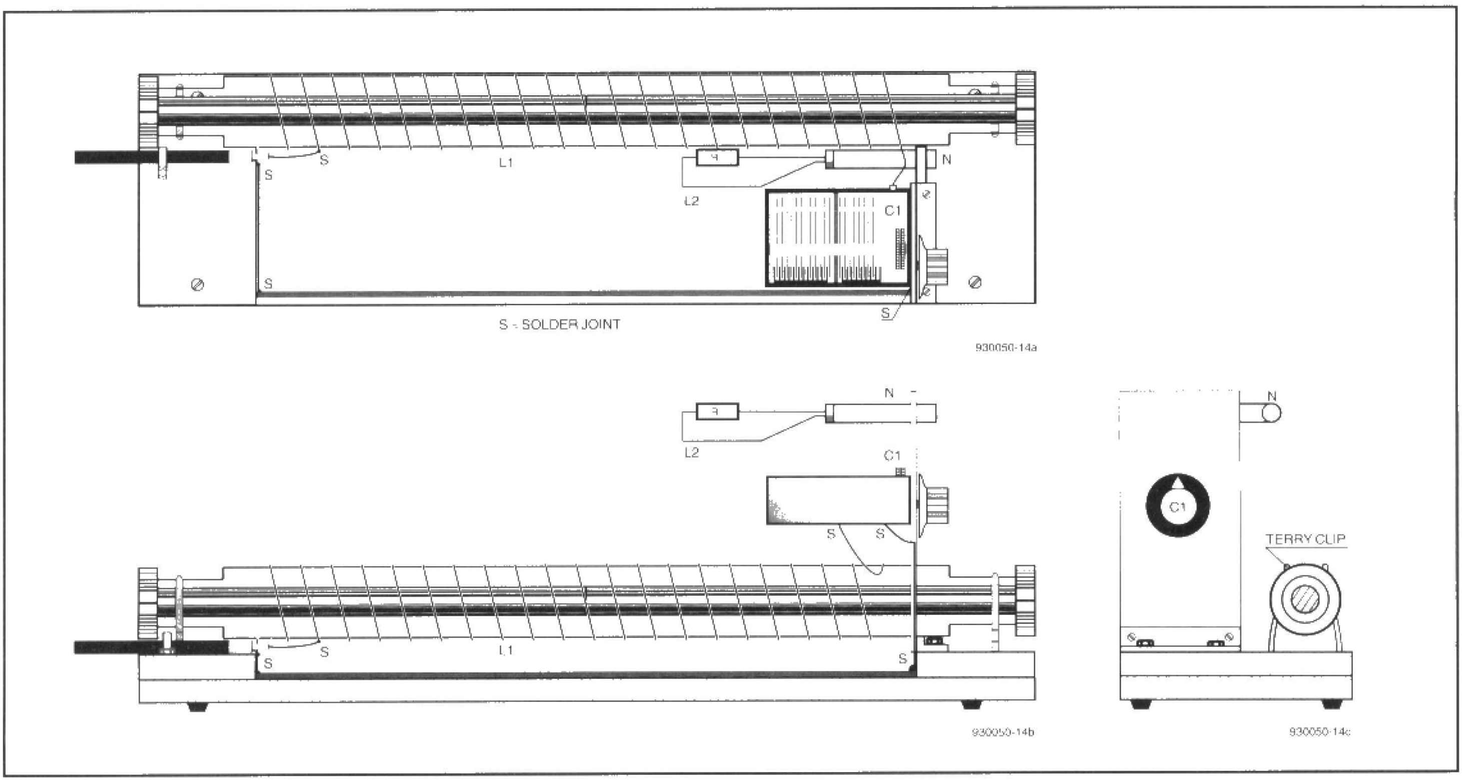


Fig. 4. Final assembly (Fig. 2 + Fig. 3. + wiring).

return. A simple radiation/resonance monitor is provided by neon lamp N. Hairpin loop L_2 is coupled loosely to L_1.

Construction

The ferrite rod is 15 in (380 mm) long and has a diameter of 0.5 in (12.5 mm). It is constructed from two 7.5 in (190 mm) long Type 61 nickel-zinc rods, securely adhered together, end to end (see Fig. 2A). One end of each rod is cleaned thoroughly by rubbing it down on fine glass paper. The two clean ends are glued together with cyanoacrylate adhesive (superglue or similar). Take great care with this operation as adhesion takes place in a few seconds; it is strongly advisable to wear old rubber (washing-up) gloves. Allow the joint to cure for several hours.

The rod is fitted into a 14 in (350 mm) long glass tube with a diameter of 2 in (50 mm)—see Fig. 2b. The prototype glass tube was a domestic rolling pin, about 14 in (350 mm) long with plastic bungs at either end. These pins, available in specialized kitchen utensil stores, are made of rugged heat-resistant glass, so that they can be filled with hot or cold water, depending on the culinary needs. If such a pin cannot be obtained, 2 in (50 mm) o/d glass-fibre tubing may be used; the end 'bungs' may be cut from soft wood about 0.5 in (12.5 mm) thick. The end bungs should be pierced at the centre and the rod ends forced through the holes so that a short length of rod protrudes at one side.

Close-wind 29 turns of SWG14 (AWG12) tinned copper wire on a $1^3/_4$ in (45 mm) diameter cardboard tube. Remove the ferrite rod from the glass tube, take the coil off the cardboard tube and force it on to the glass tube. Reduce the coil to $26^1/_2$ turns: bend the surplus wire at either end to form connecting links. Finally separate the turns evenly until the coil is expanded to 12 in (300 mm). Refit the ferrite rod.

The simple wood mounting base, shown in Fig. 3, may be teak-stained and fitted with four plastic feet as shown. Mount and bracket a 4×2 in (100×50 mm) fibre-glass board, copper-clad at one side, to the base as shown. This will be used for mounting C_1.

At the front end of the base, fasten a $^1/_8$×$^1/_8$ in (3×3 mm) brass, copper, or silver solder busbar to the wood. Solder one end of the busbar securely to the vertical fibre-glass board—see Fig. 3.

The final assembly is shown in Fig. 4. The coil assembly (Fig. 2) is attached to the base (Fig. 3) with two large Terry clips. Fit these clips to the base as far away as possible from the coil ends.

Fit C_1 as shown and solder it securely to the end of L_1. Solder the other end of the coil to the busbar. Solder a 60 in (150 cm) length of RG58 coaxial cable at a tap 2 turns from one end of L_1 (Fig. 1). Solder the screen of the cable to the busbar end of the coil. Cleat the cable to the base as shown in Fig. 4a. Every item of the assembly should be very secure and rigid. Use SWG14 (AWG12) wire for all wiring.

Mount the neon lamp as shown in Fig. 4c. It is a Type NE2 with resistor *R* in one lead, and is mounted in an insulated grommet. Secure the grommet with a U-shaped piece of wire slipped into the grommet groove and solder the wire ends to the edge of the circuit board as shown. With the neon in place in the grommet, solder its leads together and form them into a hairpin shape as shown (Fig. 4).

PARTS LIST

Qty 2 ferrite rod, Part No. R61-050-750 (Amidon Associates Inc., PO Box 956, Torrance, CA 90508, USA. Price at time of writing <autumn 1992> $18.00 each plus $6.00 airmail)
C1 = good-quality 365 pF variable capacitor, minimum 750 V, with suitable (insulated) knob.
Qty 1 glass tube—see text.
Qty 1 reel, 4 oz, of SWG14 tinned copper wire.
Neon lamp Type NE2 with integral resistor and insulated grommet (Tandy/Radio Shack).
Qty 1 busbar—see text.
Qty 2 Terry clips to fit.
Qty 4 rubber or plastic feet.
Qty 1 fibre-glass board plus angle bracket—see text.
Wood, adhesive, screws, cleats as required.

Testing & operation

Carry out initial tests with a frequency calibrated receiver and a small electronic calculator. With the antenna connected, tune the receiver to 3500 kHz. Place the calculator near the antenna so that it produces artificial noise in the receiver.

Rotate C_1 to resonance at 3500 kHz when the noise peaks. Repeat this at 4000 kHz.

Tune the transmitter to about 3550 kHz and set it up with the aid of a dummy load. The transmitter used by the writer had a 50-Ω pi-network output; an existing T-network ATU was left in circuit. Substitute the antenna for the dummy load and readjust C_1 (slightly) for maximum output on a nearby field-strength meter. Finally, adjust the hairpin loop (L_2) on the neon so that the lamp just strikes. The antenna is then ready for on-the-air tests.

The writer has used 10 watts CW output in the 80-m band between 3500 kHz and 3600 kHz. Though it will be found that core saturation will not occur until about 25 W output is used, it is **STRONGLY ADVISED IN THE INTEREST OF DOMESTIC SAFETY** not to exceed an output of 10 W.

The prototype antenna also resonates in the 7 MHz band with C_1 very near minimum capacitance, where it was found to load satisfactorily; it was, however, not air-tested.

Moreover, it also resonates on 'top band' with C_1 near maximum capacitance, but the loading has not been pursued on a transmitter.

In operation, the antenna is used on a wooden table near the operating position, so that it is just possible to rotate it slightly or tune C_1. It should be kept well away from walls and metal objects. A metal table should not be used. Connecting an earth/ground to the busbar dod not have any noticeable effect, but could be tried as individual locations and circumstances will vary from QTH to QTH. Here, it is used indoors in a first-floor room. The radiation pattern is a figure of eight with near-zero radiation off the ends. In practice, it has been found that when operating at 3560 kHz, there is a usable bandwidth of about 3552–3568 kHz, indicated by near-constant field-strength readings; coincidently, they are the frequencies at which the neon still just strikes. This method has been used as a quick, simple, satisfactory on-going check during actual operation and changing frequency. It is far quicker than using an SWR meter. It must be remembered that it is necessary to resonate C_1 to frequency, unlike the dipole where the antenna is a constant length.

Conclusion

It is hoped that this little ferrite loop transmitting antenna will encourage others to come forward and tell us about their experiences in this field, and to encourage yet others to commence experimenting. As it tunes to the 7 MH, 3.5 MHz and 1.9 MHz bands on a receiver, it seems logical that possibly it can be used on all three bands for transmission.

The writer believes that there is a future for small ferrite loop transmitting antennas, but that much work needs to be done. It is stressed that much of this article is based on personal work without any textbook backup. ■

References

[1] 'The Fe-one Experimental Transmitting Antenna' by Richard Q. Marris, *Practical Wireless*, January 1989.

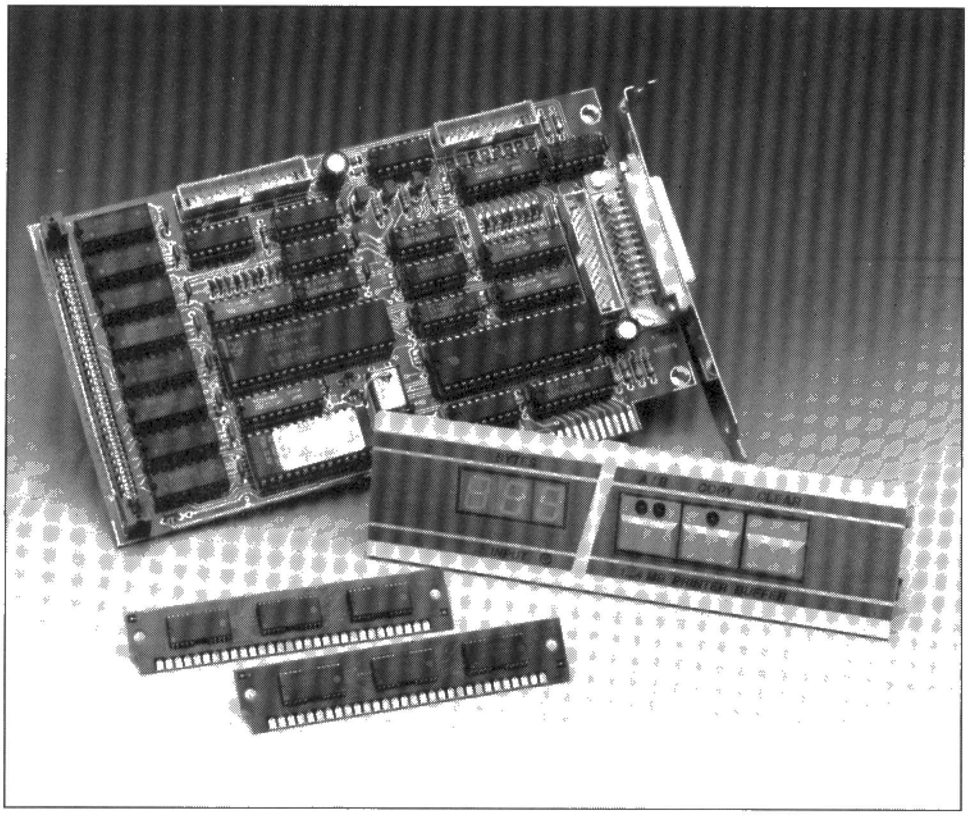
[2] 'Experimental Quadriform Ferrite Transmit/Receive Antenna' by Richard Q. Marris, *Elektor Electronics*, November 1991.

[3] *The A.R.R.L. Antenna Book*, 16th Edition, pp. 5–11.

[4] 'The Electrically Small Magnetically Loaded Multiturn Loop Antenna' by R. De Vore and P. Bohley (Ohio State University), *IEEE Transaction on Antennas and Propagation*, Vol. AP-25 No. 4, July 1977.

[5] *Iron-powder & Ferrite Coil Forms* (February 1992 issue), booklet published by Amidon Associates Inc. (USA).

4-MBYTE PRINTER BUFFER INSERTION CARD (PART 1)



Printer buffers are usually thought of as computer peripherals, that is, 'boxes' connected between the computer and the printer. Put that thought aside for a moment and you will see the advantages of the insertion card version we propose here: no power supply, no case, no extra cables. The printer buffer presented here has a pretty large capacity, while also giving you a second printer port at the same I/O address. The card allows RAM ICs as well as RAM modules (SIPP or SIMM types) to be used, and offers a total buffer capacity of 1 MByte or 4 MByte, depending on the type of memory fitted. Suitable for all types of IBM PC and compatible, be it a 286, 386 or 486 machine.

Design by R. Degen and J. Dieters

FRONT COVER PROJECT

It has been reported that the coffee consumption of today's PC user has not decreased with increased computer speeds. This phenomenon appears to be due to the fact that a fast, powerful PC is capable of producing files that take so long to print that a cup of coffee is in order. So, nothing has changed, because in the old days, the PC-XT was the 'slow' factor. To some, it is a peaceful and reassuring thought that the PC and printer are working together happily for half an hour or so. Others feel differently about this, pointing out that fast PCs should be relieved of such weary tasks as waiting for a printer. Evidently, the PC should return to its real job: computing! Alas, today's printers are and remain partly mechanical devices, which operate at snail's pace compared to the grey (mini) tower or desktop case, which is capable of crunching bits and bytes in top gear.

Unlike many other printer buffers sold by the 'box pushers' in today's computer shops, the one discussed here is 'invisible', much like a spooler program. Obviously, we are talking about a printer buffer implemented on an insertion card fitted into a free extension slot in the PC.

Prices of 1-Mbyte dynamic RAM modules have dropped considerably over the last few months, and it is safe to say that these modules are now available from almost any PC dealer. The larger 4-MByte modules are gaining ground fast, too, although their prices can still be expected to drop. This expectation allows you to fit the printer buffer with a 1-MByte memory module for the time being, and not upgrade to 4 MByte until your funds allow.

Block diagram

To begin with, it is essential to note that the present card can be used in parallel with an existing printer port. That is to say, if these two are assigned different I/O addresses. Alternatively, the printer buffer can simply replace the existing printer port. Another point worth noting is that the printer buffer has two printer outputs, which can be selected via a push-button on the front panel (see introductory photograph).

The design of the present card is derived from the earlier '4-MByte printer buffer' described in the June 1992 issue of *Elektor Electronics*. The main difference between the two buffers is that the one described here takes its input data direct from the PC extension bus, rather than from the Centronics port. If you have read and digested the earlier article, please feel free to skip the next sections and go directly to the part that deals with the construction.

The block diagram (Fig. 1) shows a dedicated system that consists of the classical components: microcontroller, memory, display, input and output. The input is formed by the IBM bus extension connector, which provides all the necessary drive signals, and, of course, the supply voltage for the printer buffer. Between the PC extension bus and the microcontroller sits a standard printer port. The microcon-

troller has only two control devices: a CLEAR button and a COPY button. It has a number of tasks, including memory, display and printer output control. Note that the output selection is controlled by a switch, which enables you to direct the print file to one of two printers.

Circuit description

Although the circuit diagram shown in Fig. 2 occupies two full pages, it can not be said to be very complex. This is mainly because all functions indicated in the block diagram are fairly easy to find back as 'real' electronics.

To the right of the diagram we find the PC extension bus connector, which has two sides, A and B. The databus lines are buffered by a bidirectional bus driver, IC22. The I/O address of the card is determined by a classic '138' based address decoder. There are three address options, which is in accordance with the number of addresses reserved for parallel printer ports (LPT1, LPT2 and LPT3) in a PC. The available addresses are $278, $3BC and $378, and their use in your PC can be checked with the aid of the DEBUG utility, as will be illustrated further on. The address selection proper on the printer buffer card is effected by fitting one of three jumpers JP3, JP4 and JP5.

Most modern PCs already have a parallel printer interface, which is usually located on an input/output card (sometimes combined with other functions). The present printer buffer is then simply added, and forms the second printer port. In any case, the existing printer port and the printer buffer should not occupy the same address. A number of address lines on the extension bus connector are decoded by IC24 and IC25. The resulting signals are applied to the 74HCT138. Two extension bus address lines, A8 and A9, are connected directly to the address decoder. If the address supplied by the PC (via the extension bus) matches the address set by the user (via the jumpers), driver IC22 is enabled, which allows data (D0 to D7) to be conveyed to the printer adapter interface (PAI), IC21. The PAI is a Type 82C11 (from UMC), which is found on many early generation PC insertion cards with a parallel interface on it. A discussion of the structure and operation of the PAI is, unfortunately, beyond the scope of this article, and interested readers are referred to the datasheets published by UMC.

At the top of the circuit diagram we find the memory of the printer buffer, IC2 to IC9, or IC20. There are three memory IC options for each version (1 or 4 MByte) of the printer buffer: (1) use 8 ICs; (2) use a SIM module (fitted in position IC20); (3) use a SIPP module (inserted into a single-in-line socket at the solder side of the board; position also IC20). For the 1-MByte version, you require either 8 off 1 Mbit (1M × 1) DRAMs (18-pin types; 256K × 4 ICs with 20 pins are not suitable), a 1 Mbyte SIMM, or a 1-MByte SIPP. For the 4-MByte version, the choice is probably limited to 4-MByte SIPP or SIMM modules, since 4 Mbit (4M × 1) ICs with 18 pins are apparently no longer produced. The memory IC shown in the circuit diagram is a 1 Mbit 18-pin type manufactured by Siemens. Any access speed equal to or faster than 100 ns will work, so that even the cheapest (slowest) versions can be used without problems. Remember, you fit **either** eight memory ICs **or** a single SIPP or SIMM module. The two can **not** be combined.

The PCAS\ line (parity column address strobe) need not bother us, since it is made permanently low. The parity bit is, therefore, not used.

It is not necessary to define the size of the buffer memory, since this is done automatically by the printer buffer control software on start-up.

Returning to our circuit diagram discussion, the microcontroller plus program memory (EPROM) are seen in the top left hand corner. The two devices are linked via an address latch, IC11. A number of octal latches Type 74HCT573 are used to interface the microcontroller to its peripherals: the PAI, the two printer outputs, the switches and the display. The relevant integrated circuits are IC1, IC14, IC15, and IC28. Two microcontroller port lines, P1.2 and P1.3, are used to drive the INPUT and COPY LEDs via invert-

MAIN SPECIFICATIONS

- **Memory size: 1 Mbyte or 4 Mbyte.**
- **DRAM memory type: DIL IC, SIPP or SIMM, any speed ≤100 ns; 3-, 8- or 9-IC types.**
- **Inexpensive memory devices.**
- **Two printer outputs (switch selectable).**
- **Low power consumption.**
- **Microprocessor-controlled (80C31).**
- **Can be used at one of three standard LPT addresses.**
- **Simple to use and 'invisible' after installation.**
- **High-quality PC insertion card.**
- **No drivers; control software in EPROM supplied with PCB.**
- **Bright LED display to follow spooling operation.**
- **Built-in DRAM test.**

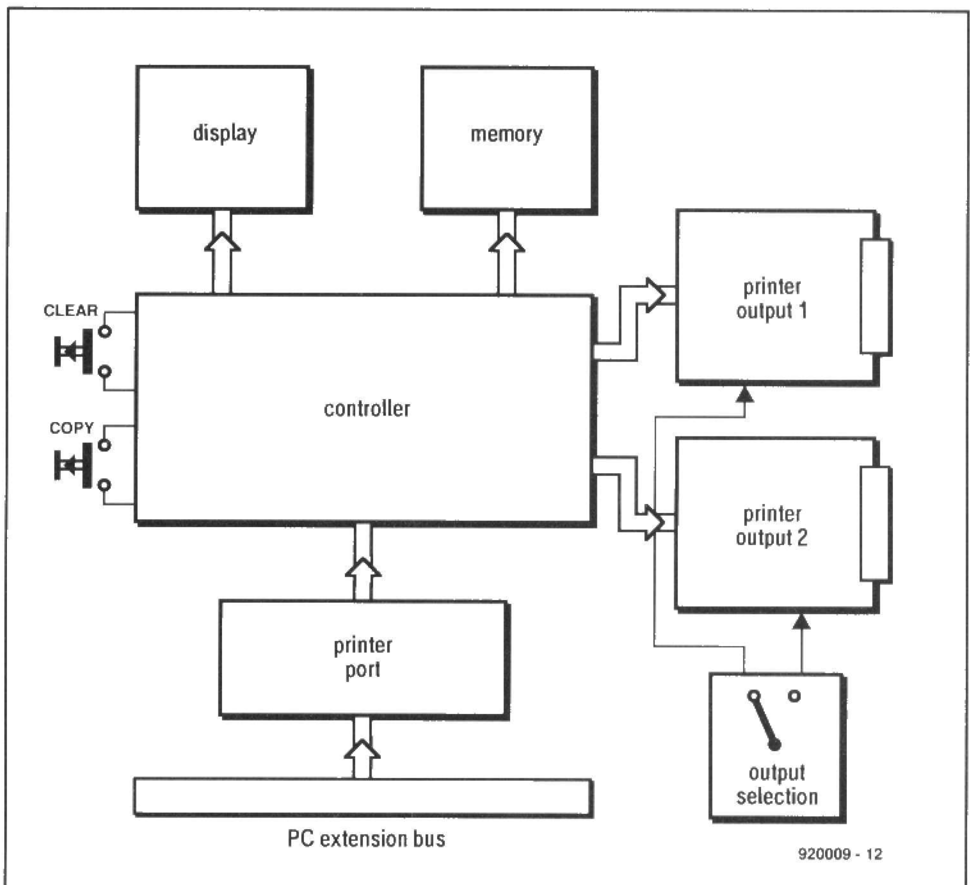


Fig. 1. Block diagram of the printer buffer insertion card.

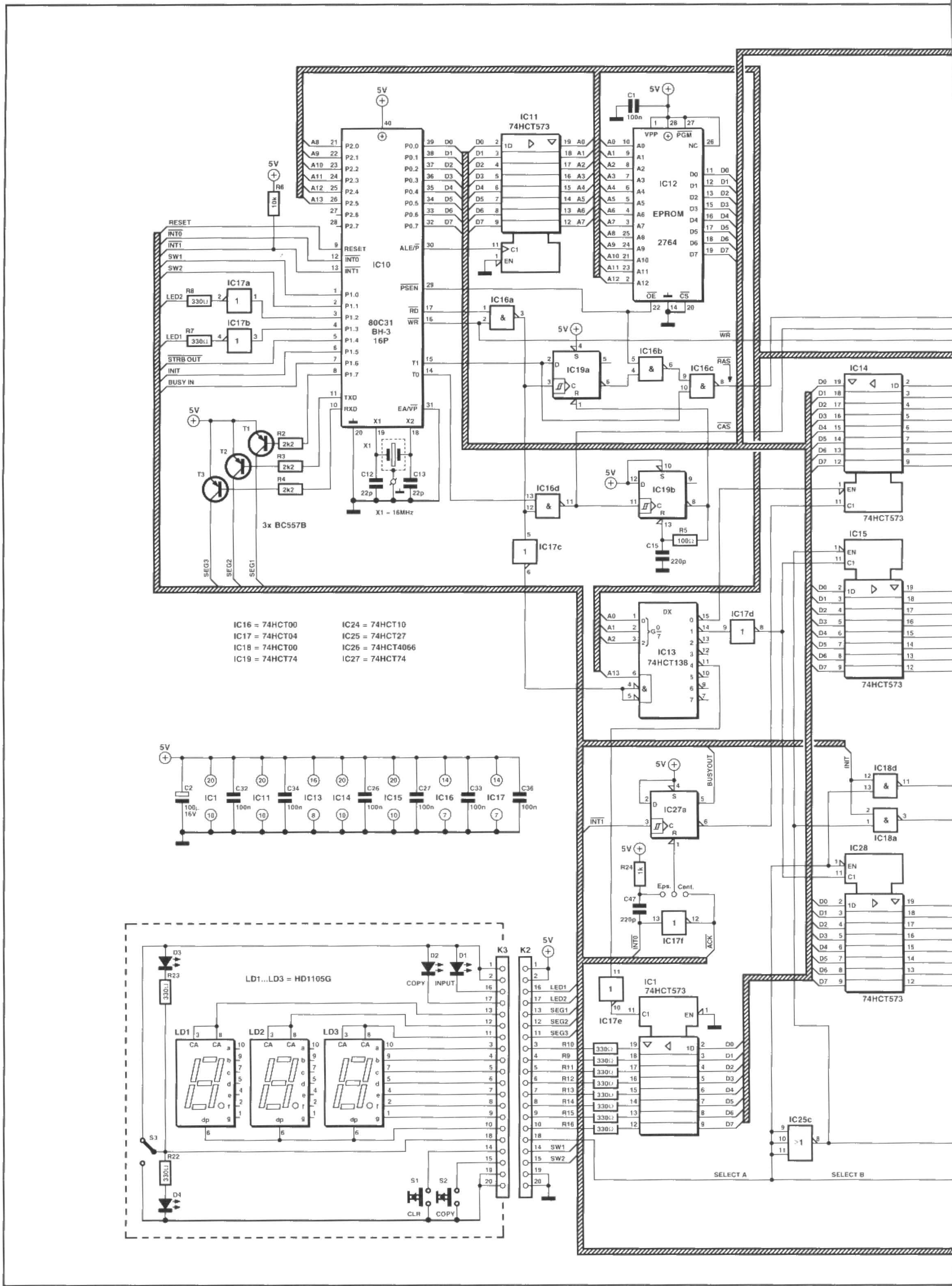
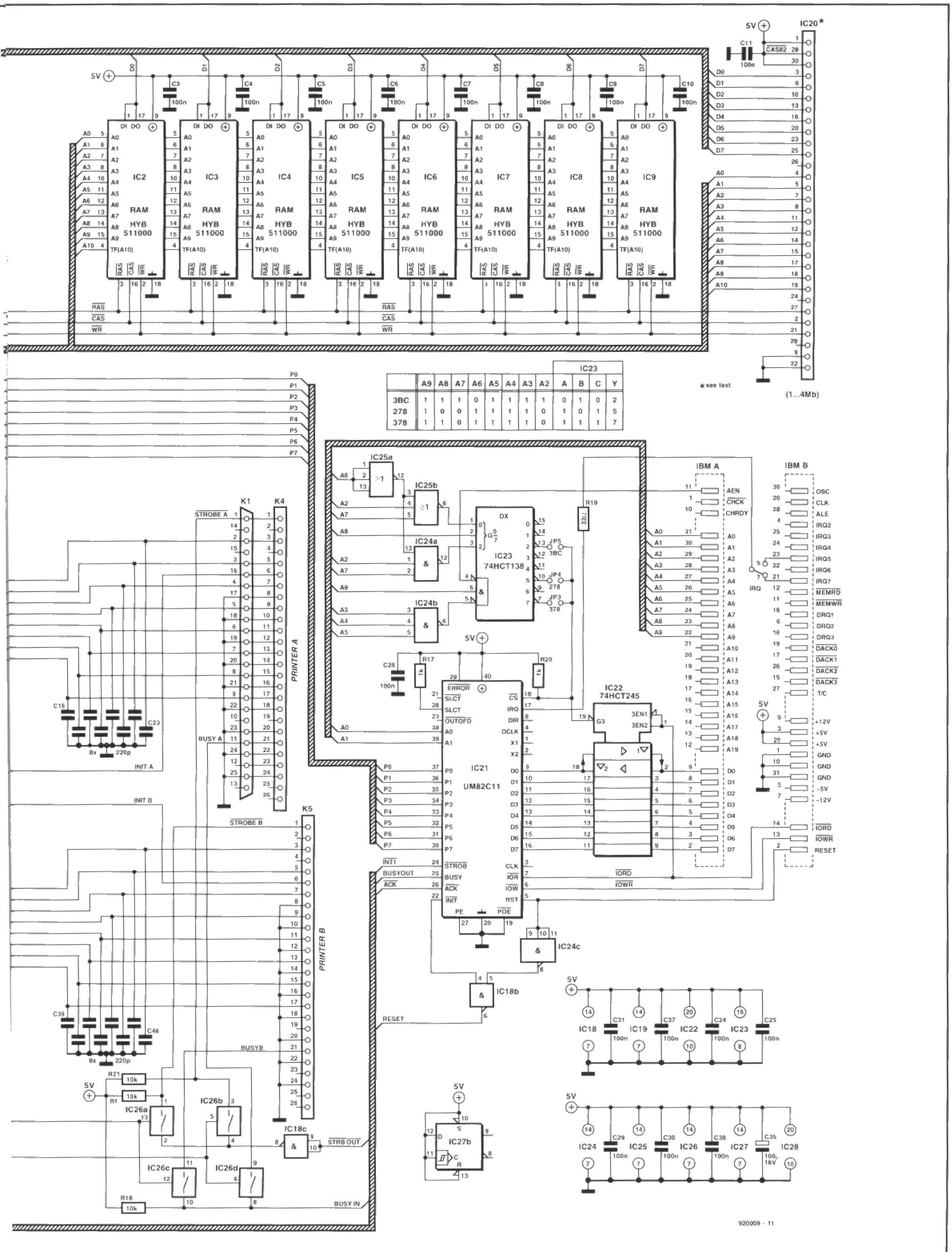


Fig. 2. Circuit diagram of te IBM PC printer buffer insertion card. The I/O address assigment options are given in the table inset below the



	A9	A8	A7	A6	A5	A4	A3	A2	IC23 A	IC23 B	IC23 C	IC23 Y
3BC	1	1	1	0	1	1	1	1	0	1	0	2
278	1	0	0	1	1	1	1	0	1	0	1	5
378	1	1	0	1	1	1	1	0	1	1	1	7

memory circuits, IC2 to IC9.

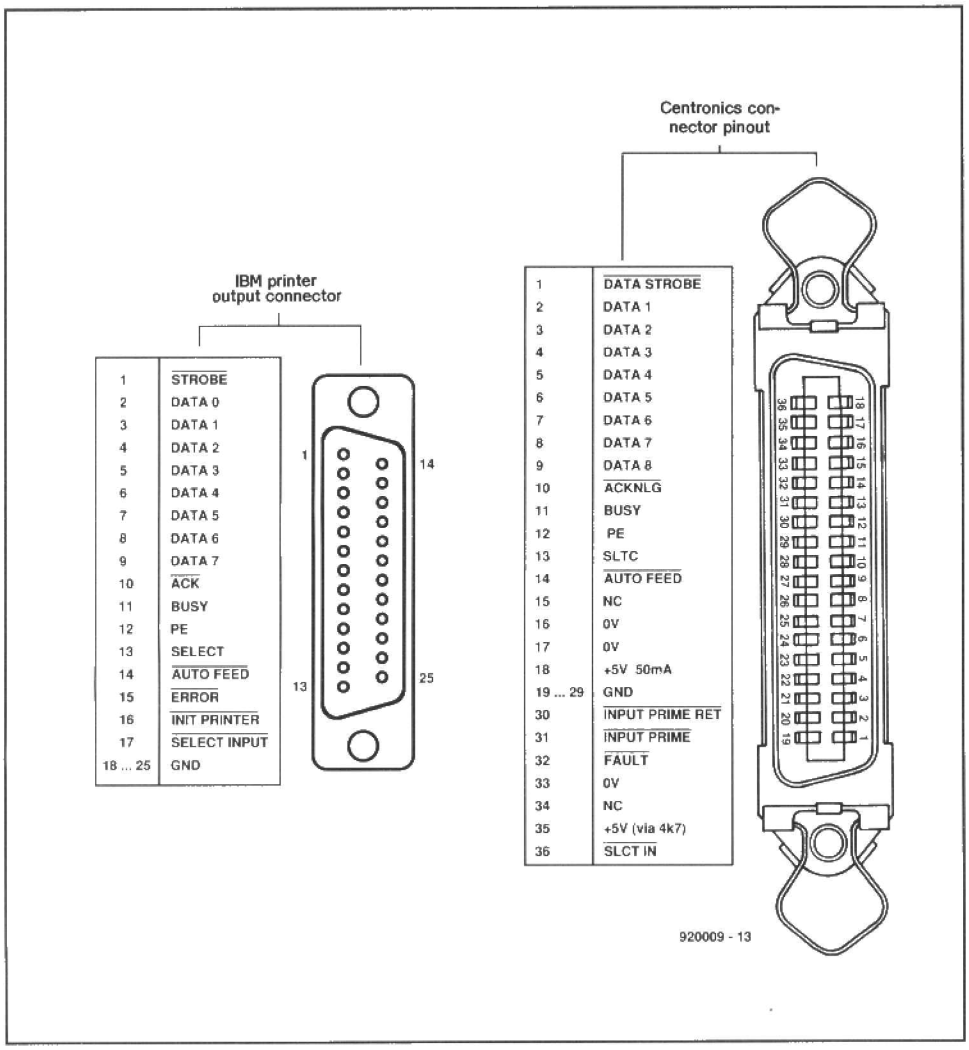


Fig. 3. For your reference: pinning of the 25-way sub-D printer connnector on the back of your PC, and that of the 36-way Centronics socket fitted on your printer.

ers. The LED display segments are under the control of three microcontroller lines: P7.1, TxD and RxD. These lines drive three transistors that furnish the necessary segment current.

The 80C31 has no on-board DRAM refresh hardware, so that this function has to be implemented in software. More about this later.

It should be noted that the printer buffer does not make use of the Acknowledge (ACK\) signal(s) transmitted by the printer(s). By default, the PC 'sees' a printer with an 'Epson' interface. This can be changed to Centronics by means of a wire jumper.

As indicated by the circuit diagram, it is possible to connect two printers, PRINTER A and PRINTER B, to the corresponding outputs of the buffer. Switch S3 ('A/B') is used to select between the two printers.

The buffer reset signal is supplied by the PC via NAND gate IC24c. After a reset, the printer is initialized.

The last block to be discussed is the display circuit, which is shown in a dashed outline in the left-hand bottom corner of the circuit diagram. This sub-circuit is connected to the main circuit via connector K3. It consists of the printer selection switch, S3, the COPY and CLEAR keys (S1 and S2), four LEDs, and three 7-segment LED displays.

The microcontroller reset signal has a very important function, which requires a short discussion. When the PC is first switched on, there is no immediate reset action on the buffer. This is because the PC runs an initialization sequence on al of its sub-functions, including the hard disk, floppy disk(s), video card, serial port, and, of course, the parallel ports. That is why a system reset has no immediate effect on the printer buffer: just wait for the initialization sequence to complete. Depending on the type of PC you have, this may take a while.

There are two possible sources for the microprocessor RESET signal:

1. The PC, whose reset signal arrives in the printer buffer via extension bus pin B2. This signal, which is active at power-on, resets the PAI, IC21, direct, while the microcontroller, IC10, is reset via gate IC18b. The same gate also conveys the PAI INIT\ output signal to the microcontroller RESET input. It should be noted that some programs, such as The Realizer, have the nasty habit of outputting a printer INITialization command. If this causes problems, take pin 4 of IC18 out of its socket or PCB hole, and force it logic high by connecting it to the +5 V supply line.

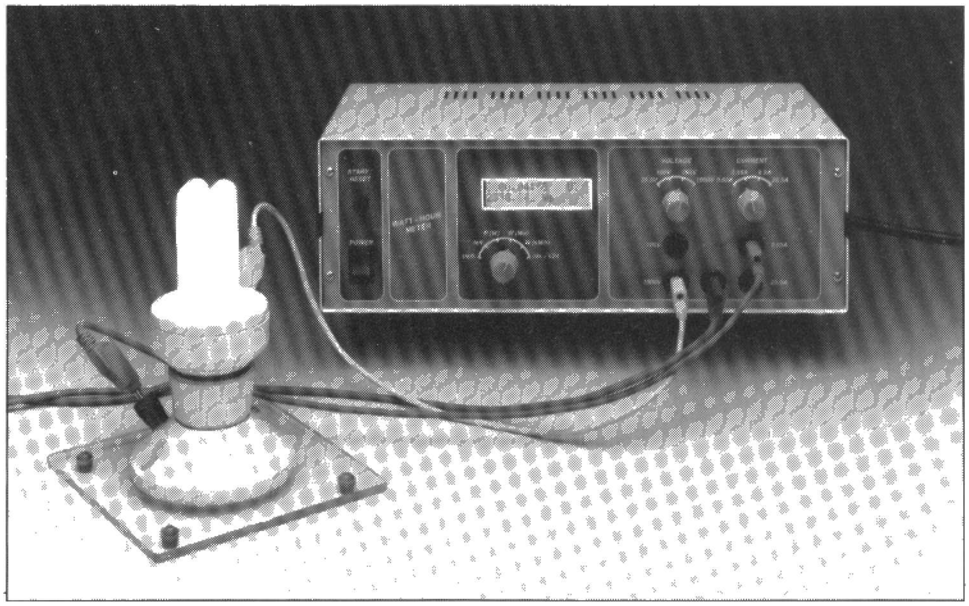
2. The printer, via the INIT signal. Note that pressing the CLEAR key does not produce a hardware reset. Depending on what the printer buffer is currently doing, either the memory is cleared (buffer in 'print' mode), or the COPY operation is aborted (buffer in 'copy' mode).

That completes the circuit description, which is purposely simplified to allow more space in this article to be given to the construction. Finally, Fig. 3 recaps the pinning of the parallel printer port connector (at the computer side) and the Centronics ('blue ribbon') socket (at the printer side). These drawings are reproduced here for reference purposes only, since it is assumed that you use a ready-made printer cable. ■

The construction of the printer buffer card will be described in next month's final instalment.

WATT-HOUR METER – PART 2 (FINAL)

Design by M. Ohsmann



Corrections. The range information given for S_2 and S_3 in Fig. 4 was, unfortunately, interchanged: S_2 controls the current ranges and S_3 the voltage ranges.

The references GND and N in Fig. 2, 4, and 7 (at the test plugs) should read 'COMMON'. The COMMON terminal is located on the front panel (lower right) between the 1000 V and 25.5 A inputs—see Fig. 11.

Finally, the caption of Fig. 7 (which is reprinted below) should begin: 'Suggested safe metering adaptor. ...'

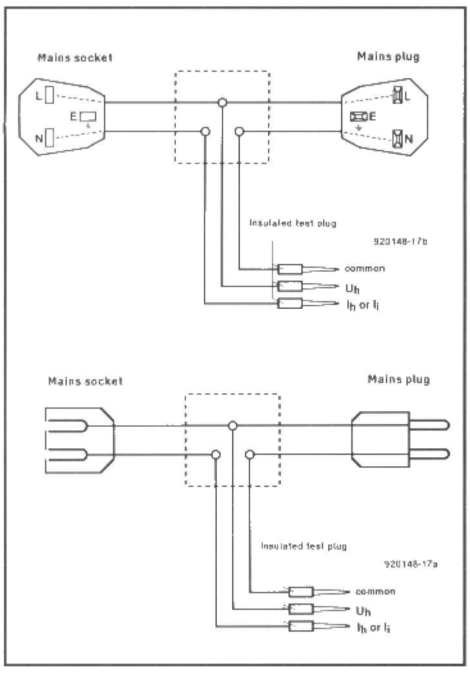


Fig. 7. Suggested safe metering adaptor...

Calibration

All calibration points are located in the two input circuits. Start the calibration by switching on the mains and ONLY THEN setting MODE switch S_4 to cal/V24. DO NOT switch on the mains with S_4 already in that position. The top line of the display should then read the measured voltage, and the bottom line, the measured current. These readings are followed in brackets by the hexadecimal value given by the A-D converter.

Presets P_2 and P_3 should still be at the centre of their travel after the initial tests. Adjust P_3 for minimum offset of IC_1. The control range of this preset should allow both a small negative value and a small positive value to be set. It is set correctly when the indicated voltage is 0 or very close to it.

Connect a voltage of 20 V and a digital multimeter (DMM) between the 100 V input and the COMMON input of the watt-hour meter (see Fig. 9). Note that the N input is the terminal at the lower right-hand side of the front panel between the 1000 V and 25.5 A inputs. Adjust P_2 until the watt-hour meter reads the same value as the DMM. This calibrates the voltage ranges.

The current ranges are calibrated in pairs, after the offset of IC_4 has been set to 0, or very nearly so, with P_7. Start with the two highest ranges, since these will have the largest deviations owing to R_{29}. The range of P_4 and P_6 has been chosen to ensure that in particular too high values of this shunt resistor can be corrected.

A current of 2–5 A is required (the closer to 6.3 A, the better). Figure 10 shows how a current of about 2 A may be obtained from a 5 V supply. Set P_4 to maximum resistance, and P_6 to minimum resistance. The amplification for the two highest current ranges is then a minimum. If the display of the watt-hour meter shows too high a value, the shunt resistor is too large. This can be remedied by taking a shorter length of wire for constructing R_{29} than stated in Part 1 or increasing the value of R_{12} to 1.5 kΩ or even 2.2 kΩ. If the watt-hour meter shows too low a current, adjust P_4 until the correct reading is obtained (as shown by the DMM). If that is not possible even with P_4 at minimum, leave it in that position and adjust P_6 until the right level of current is read. If that is still not possible the shunt resistor is too small and must be enlarged.

The two lower current ranges are calibrated with a current of about 500 mA obtained as shown in Fig. 11 (5 V and 10 Ω are, of course, also all right). Adjust P_5 until the watt-hour meter reads the same as the DMM. If that is not possible, or if the watt-hour meter responds too sharply to a small variation of P_5, increase the value of R_{11} to, say, 1.5 kΩ.

The watt-hour meter is now ready for use. However, since it is a rather different instrument than, say, a multimeter, it is advisable to read the remainder of this instalment before using it.

Using the watt-hour meter

How the watt-hour meter is to be connected to the generator (source) and the load was already shown in Fig. 2 and 6 of Part 1. The source is connected to a voltage input and COMMON, while the load is connected between a voltage input and a current input.

Setting the range is different from that of a voltmeter or ammeter, since there is no control to set the maximum value of watts or watt-seconds that needs to be measured. Since energy is to be mea-

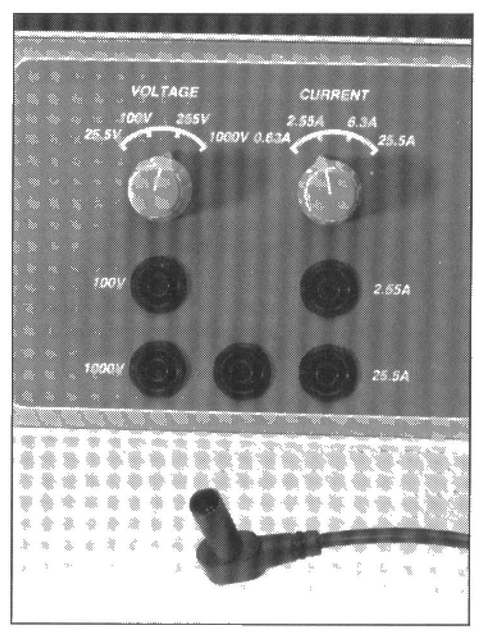


Fig. 9. The test plugs and associated input sockets must be insulated banana types.

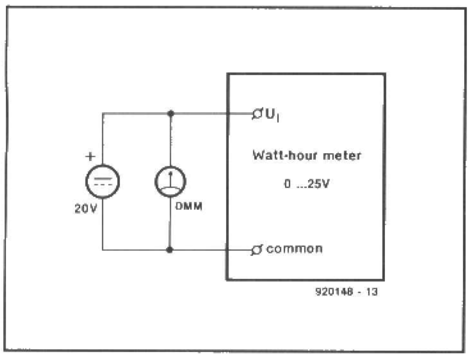


Fig. 10. Set-up for calibrating the four voltage ranges.

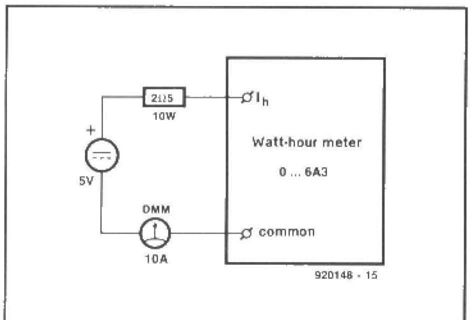


Fig. 11. Set-up for calibrating the two lowest current ranges.

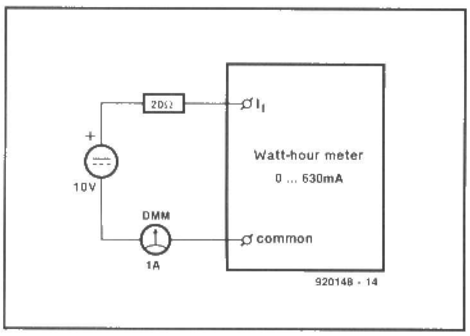


Fig. 12. Set-up for calibrating the two highest current ranges.

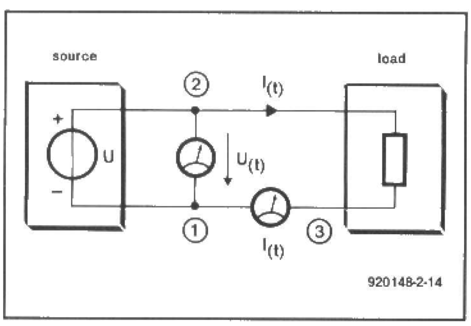


Fig. 13. Power measurement with a DC source and resistive load.

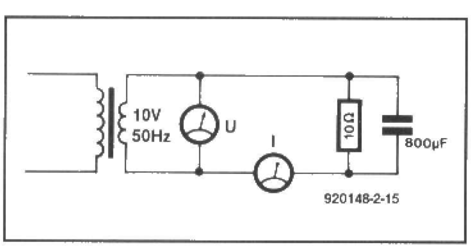


Fig. 14. Power measurement with AC source and capacitive load.

sured, both the correct voltage range and the correct current range must be selected first.

Set mode switch S_4 to position U_{eff} (in which the watt-hour meter measures voltage) and voltage selector S_3 to the required range. Note that the ranges show peak values; this means that a voltage of 240 V_{RMS} (peak value 340 V) cannot be measured in the 0–255 V range. If the range is not suitable, the display shows the message 'U?'.

Setting the correct current range (with S_4 in position I_{eff}) can be tricky. Take, for instance, a vacuum cleaner rated at 400 W operating from a 240 V mains supply. The current drawn by the appliance is then 1.6 A (RMS). However, the peak current may well be as high as 10 A. Because of this, it is wise to assume that the peak current of motors is ALWAYS higher than calculated from the motor's rating. If the range is not suitable, the display will show the message 'I?'.

Once the voltage and current ranges have been selected, set mode switch S_4 to the required position. Note that every time a different voltage range or current range is selected, the mode switch MUST be returned to position U_{eff} or I_{eff}, whatever the case may be. This must also be done whenever the display shows 'U?' or 'I?' to indicate that the range is wrong.

When alternating voltages are measured, it is wise to note the voltage and current readings on the display. Multiplying these with each other gives the **apparent power, S**, in volt-ampereas (VA). The watt-hour meter only shows the **active power, *P***, in watts (W). The difference between these powers will be reverted to later. The power factor of the load, expressed as **cosϕ**, where ϕ is the phase angle between voltage and current, is the ratio of the active to the apparent power (P/S).

When the watt-hour meter is used to measure power or energy, it registers energy, so that switching between power and energy is possible without affecting the energy registration. When, however, the mode switch is set to U_{eff} or I_{eff}, the energy measurement starts afresh: this also happens when the meter is reset with S_1 or via the RS232 interface.

Safety points

As was already stated in Part 1: **NO PART OF THE WATT-HOUR METER MUST BE CONNECTED TO THE ENCLOSURE OR (MAINS) EARTH.** Treat it as 'lethal to touch'. This is quite distinct from the electrical separation of the primary and secondary windings of the mains transformers. That isolation ensures that the watt-hour meter can be connected safely to non-mains-operated equipment.

Since the watt-hour meter is, there-

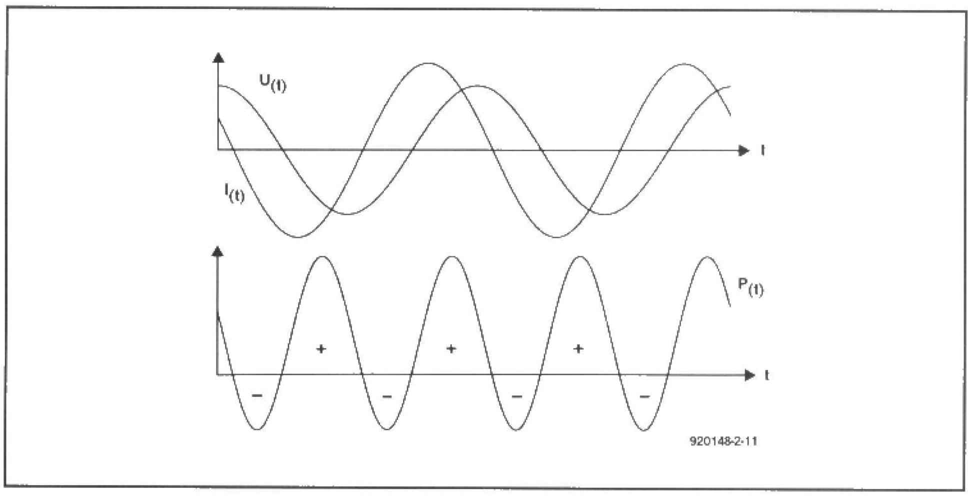


Fig. 15. Current, voltage and power as functions of time, measured as shown in Fig. 14.

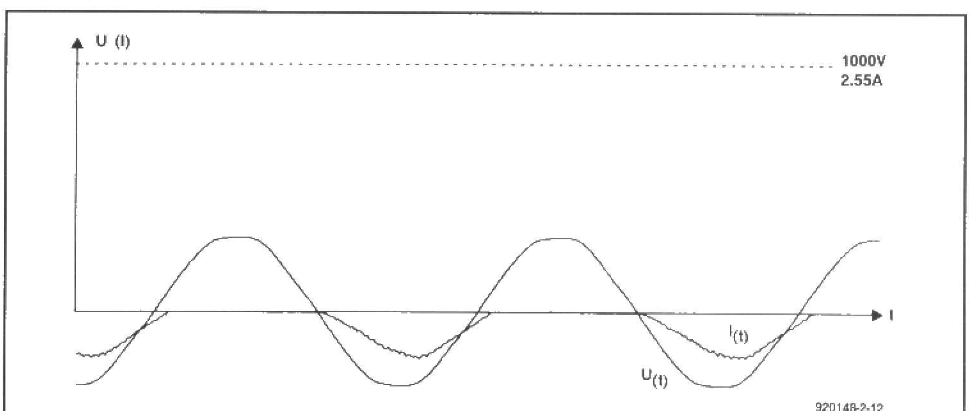


Fig. 16. Waveforms of non-sinusoidal voltage and current measured on a food mixer.

fore, at a floating potential, it may be connected to other circuits that are at whatever potential. This means, for instance, that measurements on the mains can be carried out safely. Nevertheless, it is still good practice to use an adaptor as shown in Fig. 7 (Page 26). The test plugs used MUST BE insulated types. None the less, the connection to the mains should be the last one of the operation.

Notes on measurements

It may be useful to have a closer look at the principle of measurement. In Fig. 12, the voltage across, and the current through, a resistive load are measured. It is assumed that the measuring instruments are ideal types. The power, P, dissipated in the load is obtained by multiplying the voltage and current. Assume that the generator voltage is 5 V and that a current of 0.25 A flows. The power dissipated is then 5×0.25= 1.25 W. If the polarity of the generator is reversed, the meters will show a voltage of –5 V and a current of –0.25 A. The power dissipated is then –5×–0.25=1.25 W as before. This means that the generator delivers energy to the load. In the case of resistive loads that seems self-evident. As soon as the load is capacitive or inductive, the situation is less straightforward. This is because capacitors and inductors can store and release energy (in the form of a charge or magnetic field respectively).

If, for instance, the load in Fig. 12 were a capacitor charged to 1000 V and this would be connected across a voltage source of 5 V, there would flow a negative current for a brief period, but the voltage would remain positive, provided that the source is a good one. If that negative current is multiplied with the positive voltage, the result is a negative power dissipation. This means that energy is transferred from the load to the source. The greatest effect of that role reversal comes about when it is of short duration and the source voltage is alternating, as illustrated in Fig. 13. The voltage, current and energy waveforms with respect to time are shown in Fig. 15.

Since the capacitor stores and releases charge, the resulting current is no longer in phase with the volage: it leads the voltage. At the instants that the capacitor releases energy to the transformer (and via the transformer to the mains), voltage and current are not both positive or negative, so that the power dissipation is negative.

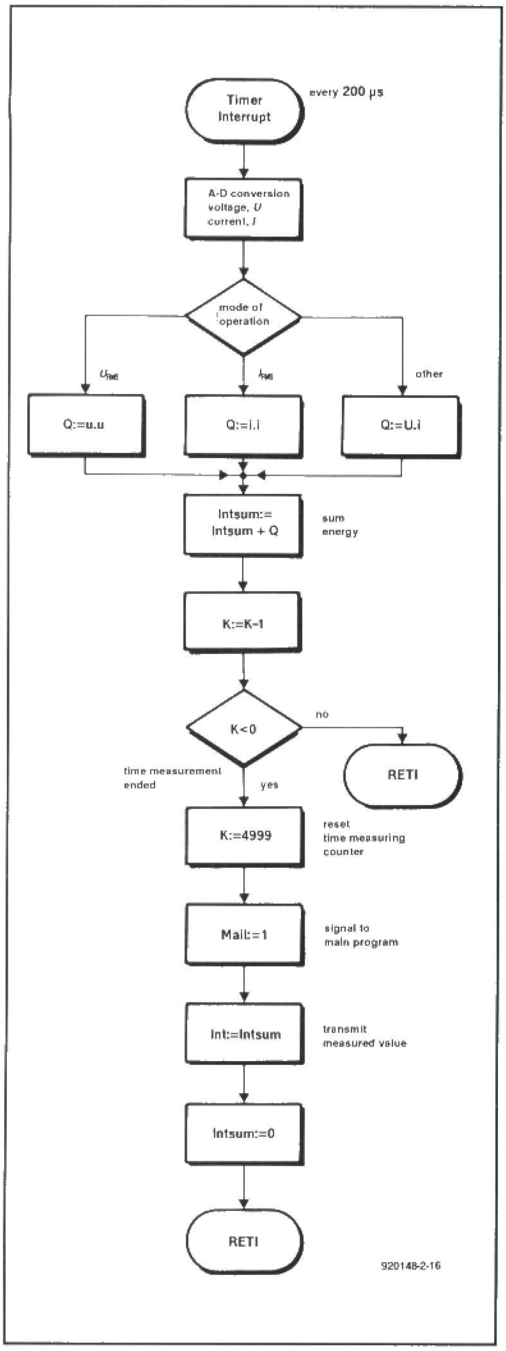


Fig. 17. Interrupt routine of the watt-hour meter.

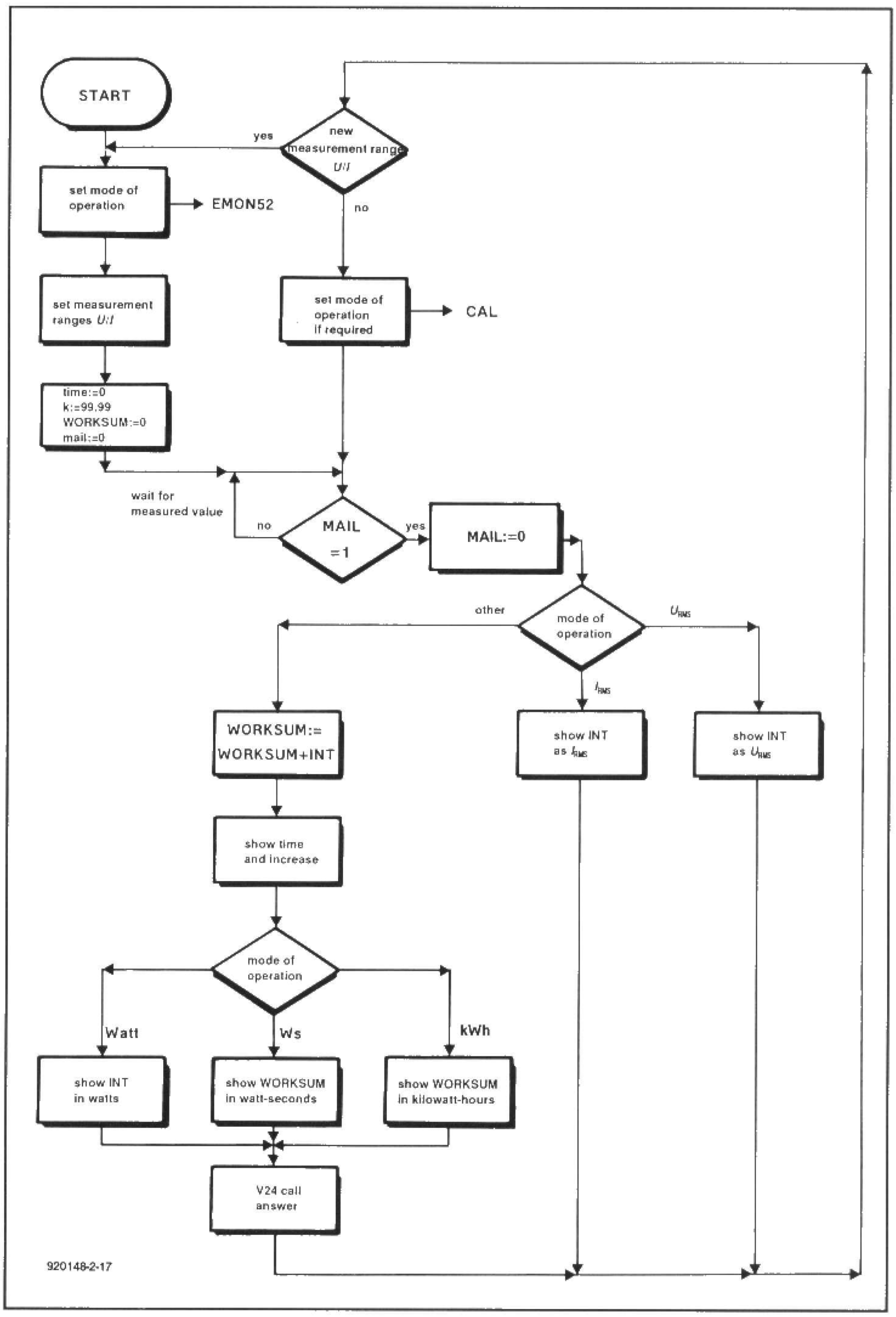


Fig. 18. The main program of the watt-hour meter.

There are three types of power possible in a circuit: (a) apparent power, S, which is the total power delivered to the load by the source; (b) active power, P, which is the power actually used by the load (and shown by the watt-hour meter): and (c) reactive power, Q, which is the power delivered to the reactive component in the circuit. In equation form:

$S=U{\cdot}I=P/\cos\phi$ [VA];

$P=U{\cdot}I{\cdot}\cos\phi$ [W];

$Q=U{\cdot}I{\cdot}\sin\phi$ [VAr].

Note that these equations are valid only for sinusoidal voltages and currents. Nowadays, there are many appliances that have a form of phase-gating, when determining the phase angle is virtually impossible as evidenced by Fig. 16. In such cases, there is no other solution than calculating the power at many instants and averaging the results over a time T, which is a whole multiple of periods. The power is then given by:

$$P=\frac{1}{T}\int_{t_1}^{t_2} p(t)\mathrm{d}t$$

Unfortunately, not everyone is capable of handling calculus; what's more, neither is a computer. The reason for this is that integration assumes an infinitely large number of instants over which the average is determined and computers cannot cope with 'infinity'. We can, however, choose a finite number of instants, if we make that number large enough for the answer to be sufficiently accurate and for the computer to be able to cope with. Here, we have chosen 5000 instants per seconds or, more correctly, 5000 samples per second. At the same time, we make the computer calculate not the power at the 5000 instants, but the energy, $W=Pt$. Substituting this in Eq. 1, we obtain:

$$W=\int_{t_1}^{t_2} p(t)\mathrm{d}t$$

This is still an integral and therefore unsuitable for the computer. However, we can approximate it as the sum of 5000 instants to obtain:

$$W=\sum_{k=1}^{5000} p(t_k)\Delta t=\sum_{k=1}^{5000} u(t_k)i(t_k)\Delta t$$

In this equation, $\mathrm{d}t$ is the time elapsed between two samples. Here, it is not really necessary to multiply 5000 times with $\mathrm{d}t$: once is enough, which saves the processor much valuable time. The equation then becomes:

$$W=\Delta t\sum_{k=1}^{5000} u(t_k)i(t_k)$$

The number of samples was chosen because the time taken by a measurement at the sampling frequency of 5000 Hz is exactly one second. This time has a number of advantages. To start with, in practice it is invariably a multiple of the period, which is a prerequisite for the computation. Also, it is long enough to ensure that any deviations are negligible. Moreover, it makes the task of the computer simpler since power is calculated by dividing energy by the time in seconds, so that here it is divided by 1.

Software

The most important task of the software is the gathering, 5000 times a second, of the measurand (that is, the measured values). This is done with the aid of a timer (in the controller) that gives an interrupt every 200 μs. At each interrupt, the controller stops all other operations and carries out a measurement.

The flow diagram of the interrupt routine is given in Fig. 17. First, the voltage and current are measured via the A-D converters in the controller. The next operation depends on the measurement to be carried out. If it is the RMS value of voltage or current, the measurand is multiplied by itself; if it is power or energy, voltage is multiplied by current. This intermediate result is summed during one second (that is, 5000 times) in the variable «intsum». Then, counter «K» is reduced by 1, whereupon a check is made whether 5000 measurands have been gathered in «intsum» (or whether the measurement duration was 1s). If this is not so, the processor is redirected to the main program to carry out other tasks. If 5000 measurands have been gathered, the interrupt routine first prepares the result for the main program in variable «int», reset variables «K» and «intsum», and signals to the main program via variable «mail» that new data are available. After that, the interrupt routine is ready for the next series of 5000 measurands and the processor is redirected to the main program until the next interrupt.

The flow diagram of the main program is shown in Fig. 18. In the absence of other tasks, the processor is at standby at «MAIL=1» until a signal from the interrupt routine indicates that there is new data in variable «int». The data is then processed and made legible on the display.

The most intensive work is carried out when the RMS value of voltage or cur-

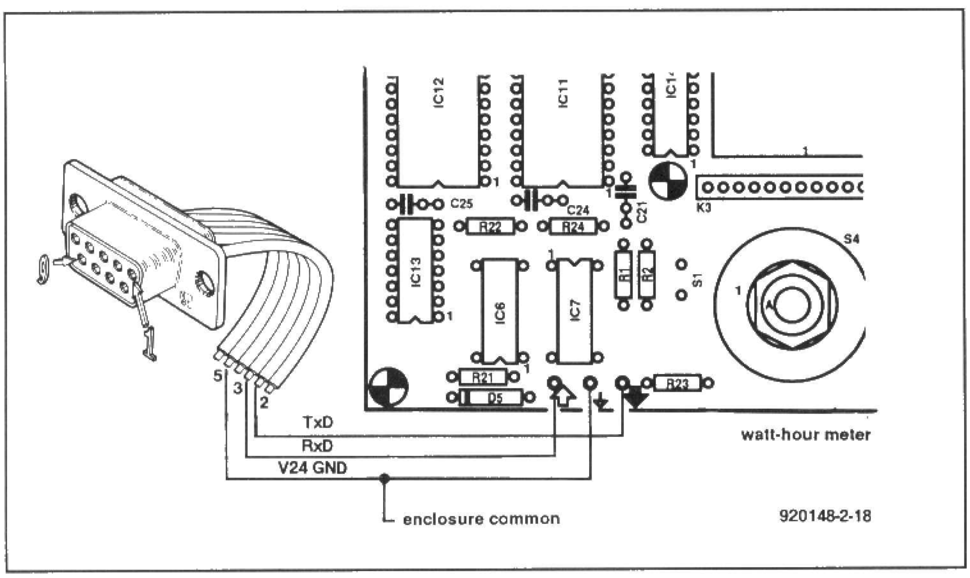


Fig. 19. The RS232 (V24) interface.

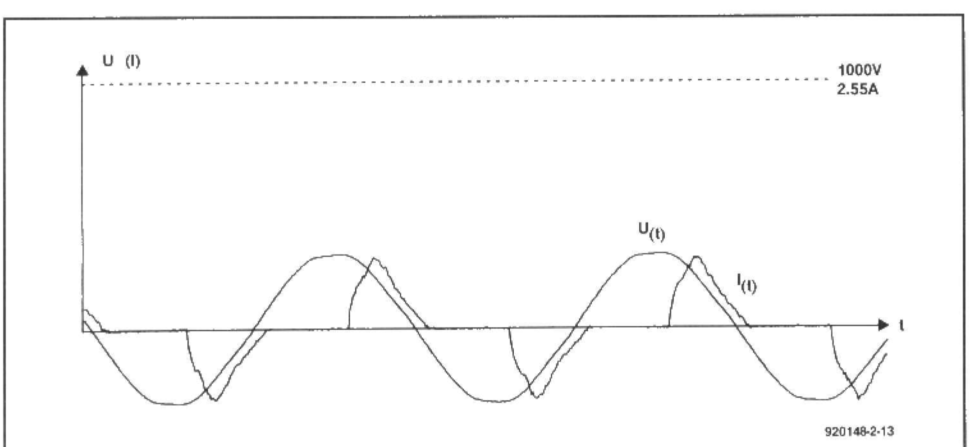


Fig. 20. Waveforms of current and voltage of a vacuum cleaner with phase gating.

rent is to be displayed. To do this, the processor must calculate the root of the value in «int». This is a very simple operation on a pocket calculator, but in assembler program it takes about one sheet. The processor works with this equation:

$$U_{\mathrm{RMS}} = \left[\Delta t \sum_{k=1}^{5000} u^2(t_k) \right]^{1/2}$$

$$I_{\mathrm{RMS}} = \left[\Delta t \sum_{k=1}^{5000} i^2(t_k) \right]^{1/2}$$

Once this computation has been completed, the binary measurand is converted into a decimal value and displayed. After that, the processor checks whether any controls have been altered. If not, it returns to standby.

When the watt-hour meter is used for power or energy measurements, elaborate computations, other than converting the measurand into a decimal value, are not required. The only operation is adding (or subtracting if energy has been released) «int» to the measured value of energy already in «worksum». The time elapsed since the energy measurement began must be increased by 1s. The time is displayed as hours;minutes;seconds.

Finally, after all the measurement data has been processed, the processor checks whether information has been requested via the RS232 interface.

RS232 interface

When the watt-hour meter measures power or energy, the measurement data are available for a computer via the serial interface—see Fig. 19. All that is needed is a suitable command to the watt-hour meter at 4800 bit s^{-1}. The available instructions are:

R=reset the energy measurement;
P=display power;
W=display energy in watt-seconds;
K=display energy in kilowatt-hours;
U=display RMS value of voltage;
I=display RMS value of current;
T=show measurement time;
V=show EPROM version.

At instructions U and I, the summing of energy is interrupted for 1s because voltage and current are measured independently. This should be done, therefore, only if the effect of the interruption on the energy measurement is irrelevant or negligible.

Note also that the response to an instruction may take up to 0.2s.

The answer to an instruction is given as a line that is always terminated by OD_{HEX}/OA_{HEX}(CR/LF).

Programming the watt-hour meter

The watt-hour meter is accessible as a freely programmable controller system via integral monitor program EMON52. When mode switch S_4 is in position cal./V24, and the watt-hour meter is reset, this program will start. The assembler and communication software of the '8051-8032 course'* may also be used. Furthermore, that course may be used to become acquainted with the assembler language of the controller in the watt-hour meter.

The waveforms in Fig. 20 were obtained from data sent to a computer by a routine that measured voltage and current 1000 times at a sampling frequency of 10 kHz. Note that the bandwidth was restricted to 5 kHz. ■

* *Elektor Electronics*, February–November, 1992.

FIGURING IT OUT

PART 3 – SIMPLIFYING NETWORK ANALYSIS

By Owen Bishop

This series is intended to help you with the quantitative aspects of electronic design: predicting currents, voltage, waveforms, and other aspects of the behaviour of circuits. Our aim is to provide more than just a collection of rule-of-thumb formulas. We will explain the underlying electronic theory and, whenever appropriate, render some insights into the mathematics involved.

To begin with, we look at a routine based on the **superposition law**. This law applies only to circuits composed of resistances and powered by ideal current sources and ideal voltage sources. There are many circuits that can be considered to be of this type, provided that the energy sources do not depart too far from the ideal behaviour, so this law is useful over a wide range of problems.

Answers to Test Yourself – Part 2

1. 1.725 C; 34.5 V

2. $Q=\int_{0.8}^{1.0}(0.5+3t^2)\mathrm{d}t=(0.5+t^3)_{0.8}^{1.0}=0.588$ C

3. $Q=\int_{0}^{T}-8\sin 4t\ \mathrm{d}t=(2\cos 4t)_0^T=-2.45$ C,

where T is the time in which the charge is reduced by 2.45 C to become zero. $4T=\cos^{-1}-0.225$, and $T=0.45$ s.

Superposition law

Briefly, the law states that, if there are a number of current and voltage sources in a circuit, each of these contributes **independently** to the currents flowing or to the voltages developed. The action of each source is **superposed** on that of the other sources. As a result of this law we can simplify the analysis of the circuit by assuming that one or more of the sources has been put out of action or eliminated. This leaves us free to consider the effects of the remaining active sources, usually one at a time.

A circuit may have one or more sources, which may be voltage sources, current sources, or both. A current source is eliminated by re-connecting it (in the imagination) so that no current can flow from it. We simply break the circuit at that point and replace the current source with an **open circuit**. By contrast, a voltage source is put out of action by connecting a wire across its terminals, so that no potential difference can develop. We replace it with a **short circuit**.

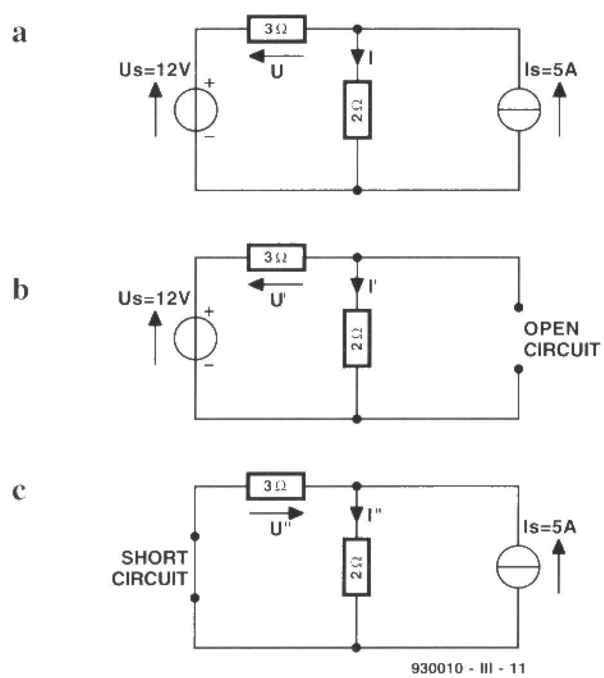


Fig. 21. Illustrating the superposition technique.

Consider the example in Fig. 21a. The circuit has one voltage source, U_s=12 V, and one current source, I_s=5 A. The problem is to calculate the voltage and current shown as U and I in the figure.

Step 1. Replace the current source with an open circuit (Fig. 21b). The circuit now consists of two resistors in series with the voltage source. The pd, U', across the 3 Ω resistor caused by the voltage source is

U'=3/(3+2)×12=7.2 V.

The polarity of this pd is indicated by the arrow in Fig. 21b.

Current I' through the resistors caused by the the voltage source is

I'=12/(3+2)=2.4 A.

There is also a current I'' caused by the current source but, for the moment, we are ignoring this.

Step 2. Restore the current source but remove the voltage source, replacing it with a short circuit (Fig. 21c). Now the circuit consists of two resistors in parallel across the current source. Voltage U'' across these caused by the current source is

U''=–2×3/(2+3)×5=–6 V.

The negative sign indicates that the voltage is opposed to U' developed by the voltage source. Current I'' is calculated from the current division rule:

I''=(3/5)×5=3 A.

The current flows in the same direction as I', so both have a positive sign.

Step 3. Superpose (sum) the voltages and currents, taking account of sign:

U=U'+U''=7.2+(–6)=1.2 V;

I=I'+I''=2.4+3=5.4 A.

Another example appears in Fig. 22a, where there are two current sources.

Step 1. Eliminating source 1 gives the circuit of Fig. 22b. This

has two 1 Ω resistors inseries, shunted by a 2 Ω resistor. This is equivalent to 2 Ω in parallel with 2 Ω, so current I_{s2} divides equally so that I'=3 A.

Step 2. Eliminate source 2, giving 1 Ω in parallel with 3 Ω. Current I_{s1} divides in the ratio 1:3 with the larger current, I'', flowing through the 1 Ω resistor. Thus,

$I''=(3/4)\times8=6$ A.

By superposition:

$I=I'+I''=3+6=9$ A.

Given this result, it is possible to use KCL (see Part 1) to calculate all the other currents in the network (Fig. 23a). We then use Ohm's law to calculate the voltages (Fig. 23b).

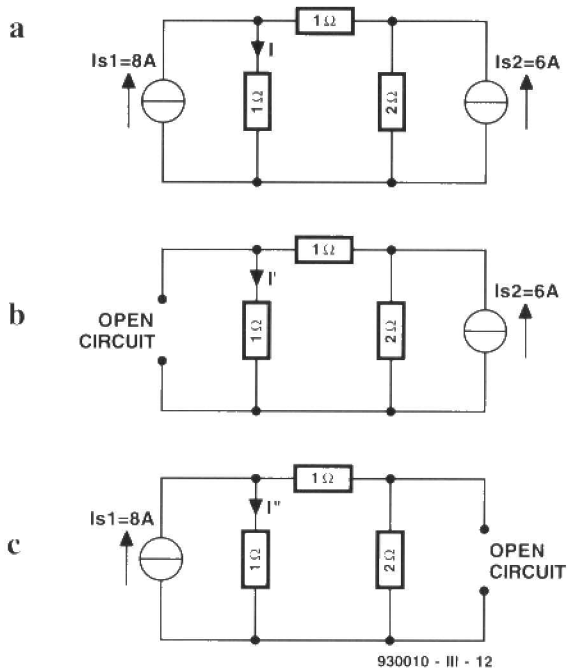


Fig. 22. Another example of superposition.

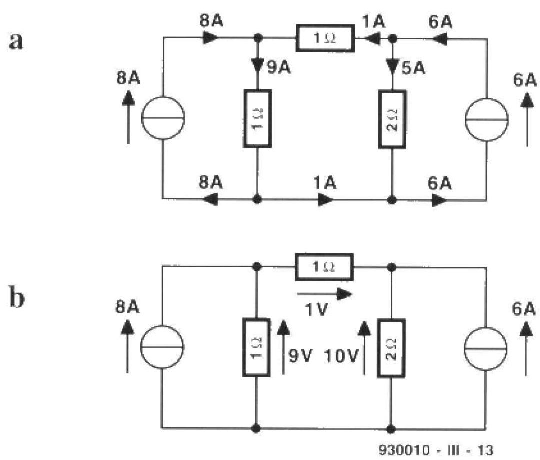


Fig. 23. Complete analysis of Fig. 2a: (a) currents; (b) voltages.

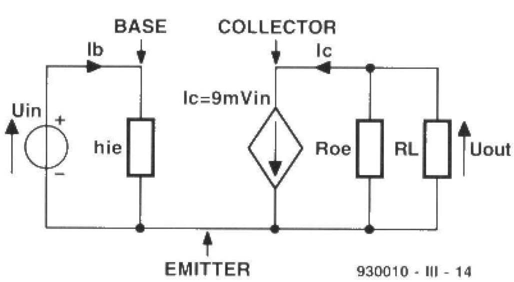


Fig. 24. Simplified equivalent circuit of a common-emitter amplifier.

Ideal sources

It is important to remember that the superposition law applies only to **ideal** sources. An ideal voltage source maintains a specified pd across its terminals, no matter what current is drawn from it. The specified pd may be constant, vary steadily with time (a ramp), be pulsed, or alternate periodically. Varying pds make the calculations more complicated, but they do not alter the application of the law.

Unfortunately, few practical voltage sources are ideal; almost all have some internal resistance, or output impedance. The precision of the analysis depends on the magnitude of this. If at all feasible, we assume that departures from ideal behaviour are small and can be ignored. If it is known that they are too large to be ignored, we can allow for them to a certain extent by adding to the circuit diagram a resistance of suitable value in series with the source.

Similar remarks apply in the case of an ideal current source.

Controlled sources

In certain circuits there may be voltage or current sources whose output depends on a current or pd existing in another part of the same circuit. An example is the collector current of a bipolar transistor, which is controlled by the base current. The equivalent circuit of a common-emitter amplifier (Fig. 24) shows the collector current as the product of the input voltage, U_{in}, at the base and the transconductance, g_m. Resistor R_{oe}, of value $1/h_{oe}$ (h_{oe} is the common-emitter output conductance), represents the the effect of the slope of the collector characteristic. It corrects for the fact that this slope means that the controlled current source is not ideal, and is an example of how we may introduce hypothetical components into a circuit in order to compensate for recognized imperfections.

Controlled sources must not be eliminated when a network is being analysed by superposition. Consider the circuit of Fig. 25a, which includes a controlled voltage source. The voltage produced by this source is controlled by I_1, the amount of current flowing through resistor R_1. In this example, the voltage equals $4I_1$ and must be added into the equations at all stages.

Step 1. Eliminate the current source, but leave U_1 and the controlled source active—see Fig. 25b. The current now flowing is denoted by I_1'. In calculating the voltage from the controlled source we use this value. From KVL (see Part 1) we find:

$-3I_1'-4I_1'=14;$

$\therefore\ -7I_1'=14$, so that $I_1'=2$ A.

Step 2. Eliminate the voltage source, but leave the current source and the controlled voltage source active. Indicating current and voltage at this stage by a double prime, and using KCL at node A—see Fig. 26c:

$I=3.5=-I_1''+I_2''=$

$=U''/2+(U''-4I'')/1=$

$=U''/2+U''-4(-U''/2).$

$\therefore\ 7U''/2=3.5$, so that $U''=1$ V.

From this we find

$I_1''=-U''/2=0.5$ A.

By superposition:

$I_1=I'+I''=2+(-0.5)=1.5$ A.

This example shows how KVL and KCL are used to solve the problem when controlled sources are present.

A further example

The network of Fig. 26 does not contain any current sources, but there are external sources supplying 2 A and 1.5 A to nodes A and B respectively. As an extension of KCL, the total current leaving any section of a network equals the total current entering it. Thus, there must be 3.5 A leaving the network at node D. The problem is to find the current passing along branch AB.

We apply the superposition

technique by eliminating each entering current in turn. With the 1.5 A current eliminated, by disconnecting the network from the external circuit at B, 2 A flows in at A and out at D. The network consists of 15 Ω in parallel with 15 Ω, so the current divides equally at A, and thus I' is 1 A, flowing from A to B. Eliminating the 2 A current means that 1.5 A enters at B and leaves at D. Now the network consists of 25 Ω in parallel with 5 Ω. One fifth of the current passes along BCD. Thus, I'' is 0.3 A flowing from B to A, or –0.3 A from A to B. By superposition:

$$I=I'+I''=1+(-0.3)=0.7\ \text{A}.$$

Thevenin's theorem

This theorem states that any 2-terminal network consisting of one or more voltage or current sources and one or more resistances may be represented by a single voltage source in series with a single resistance (Fig. 27). If various external circuits are connected to the two terminals, the behaviour of the complex network and the Thevenin equivalent circuit are indistinguishable. Replacing a complex network by its Thevenin equivalent considerably simplifies subsequent calculations.

The essential features of the Thevenin equivalent are

- the Thevenin voltage, U_{Th}, and
- the value of the Thevenin series resistor, R_{Th}.

Methods of calculating U_{Th} and R_{Th} are as follows.

Thevenin voltage. Given the Thevenin equivalent circuit on the right of Fig. 27, let us suppose that terminals A and B are not connected to any external circuit. In this condition, no current flows through the resistor and there is no voltage drop across it. The open-circuit voltage across AB equals U_{Th}. Correspondingly, given a network, we find U_{Th} for its Thevenin equivalent by calculating the open-circuit voltage of the network. We can use the superposition method or other methods to do this.

Thevenin resistance. Imagine that terminals A and B of the Thevenin equivalent are short-circuited. A short-circuit current, I_{sc}, flows from A to B. The magnitude of this current is given by:

$$I_{sc}=U_{Th}/R_{Th},$$

from which we obtain the Thevenin resistance:

$$R_{Th}=U_{Th}/I_{sc}.$$

Since U_{Th} equals the open-circuit voltage, U_{oc},

$$R_{Th}=U_{oc}/I_{sc}.$$

We have already calculated U_{oc}, so all we now have to do is to calculate I_{sc} and divide as shown.

The example in Fig. 28 demonstrates the procedure. The open-circuit voltage is the pd across the voltage source and the 2 Ω resistor. Note that the 3 Ω resistor does not come into the calculation because there is no current flowing along it when AB is an open circuit and, therefore, no pd develops across it. Ignoring this resistor, the circuit consists of 10 Ω in series with the source. There is a drop of 1 V across the 2 Ω resistor, and a rise of 5 V across the source. The open-circuit voltage is 4 V, so that U_{Th}=4 V.

The basis of calculating the short-circuit current is clearer if we re-draw the circuit, including

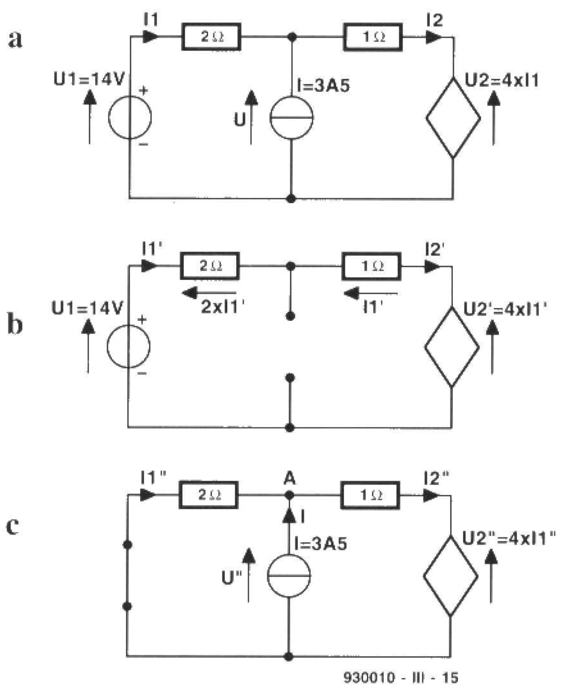


Fig. 25. A circuit with a controlled voltage source.

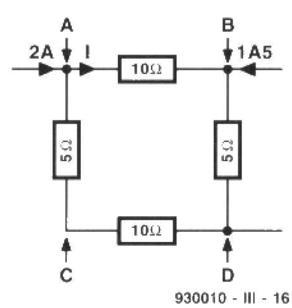


Fig. 26. Superimposed currents.

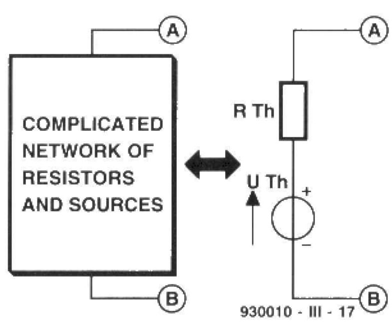


Fig. 27. Thevenin equivalent circuit.

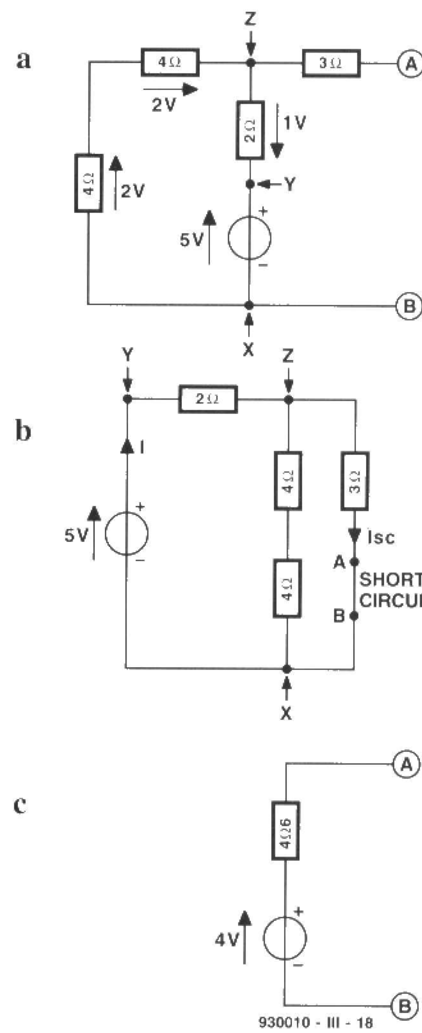


Fig. 28. Finding a Thevenin equivalent circuit

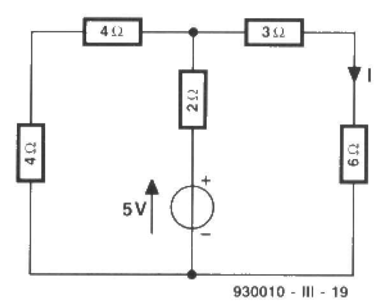


Fig. 29. Applying the Thevenin equivalent method.

the short-circuit link AB—see Fig. 28b. The equivalent resistance of the four resistors is 46/11 Ω. Current I is

$$I=5\times11/46=55/46\text{ A}.$$

By the rule of current division, we find:

$$I_{sc}=55/46\times8/11=20/23\text{ A}.$$

Finally, we calculate R_{Th}:

$$R_{Th}=U_{oc}/I_{sc}=4\times23/20=4.6\ \Omega.$$

Figure 28c shows the Thevenin equivalent resulting from the above calculations.

Alternative method for R_{Th}

Reverting to Fig. 25a, replace the voltage source by a short circuit, and then find the equivalent resistance of the circuit by reduction. A current source would be replaced by an open circuit. The two 4 Ω resistors reduce to one 8 Ω resistor in parallel with 2 Ω, which reduces to 1.6 Ω. The 3 Ω resistor is in series with this, making a toal Thevenin resistance of 4.6 Ω as obtained earlier.

Using the Thevenin equivalent

The most obvious advantage of the equivalent is that it makes it very easy to predict how such a simple circuit will behave when connected to an external circuit. Having calculated how the equivalent will behave, we know that the original complicated network will behave in exactly the same way.

An application of the method is illustrated by the network of Fig. 29a. It is required to find current I. To shorten the explanation, we use the same network as in Fig. 28a with a 6 Ω resistor connected across AB. The first step in the calculation is to remove the 6 Ω resistor. Then find the Thevenin equivalent of the network remaining. This we did in the previous section with the result shown in Fig. 28c. Finally, attach the 6 Ω resistor to the Thevenin equivalent at A and B and calculate I:

$$I=U_{Th}/(R_{Th}+6)=$$

$$=4/(4.6+6)=0.38\text{ A}.$$

Summing up this method: to find a voltage or current in a branch of a network, 'remove' the resistor, reduce the rest of the network to its Thevenin equivalent, 'connect' the resistor to this and calculate the voltage or current.

The method is applicable to more intricate networks. To find I in Fig. 30a, consider the left and right sides of the network separately, ignoring the branch AB in each case. It is left to the reader to confirm that the equivalent of the left side is specified by U_{Th}=6 V and R_{Th}=3 Ω. The equivalent of the right side has U_{Th}=4 V and R_{Th}=1.8 Ω. We now 'reassemble' the network, but connect the **equivalents** to branch AB—see Fig. 30b. Using KCL to calculate voltage U at node A relative to node B which we consider to be at 0 V:

$$(6-U)/3+(4-U)/1.8+U/3=0,$$

whence U=7.6 V.

From this result, we find that

$$I=7.6/3=2.53\text{ A}.$$

The final example—see Fig. 31a—has a controlled source, but no independent source. Since the voltage from the controlled source depends on current I, which is 0 when AB is an open circuit, U_{Th}=0. To calculate R_{Th}, we apply an arbitrary voltage to AB. For convenience, make this voltage equal to 1 V. Applying KCL to the currents flowing in and out of node C:

$$I=1/5+(1-4I)/2\text{ A}.$$

The last term on the right expresses the fact that the pd across the 2 Ω resistor is the difference between the 1 V applied externally and the $4I$ V

a

b

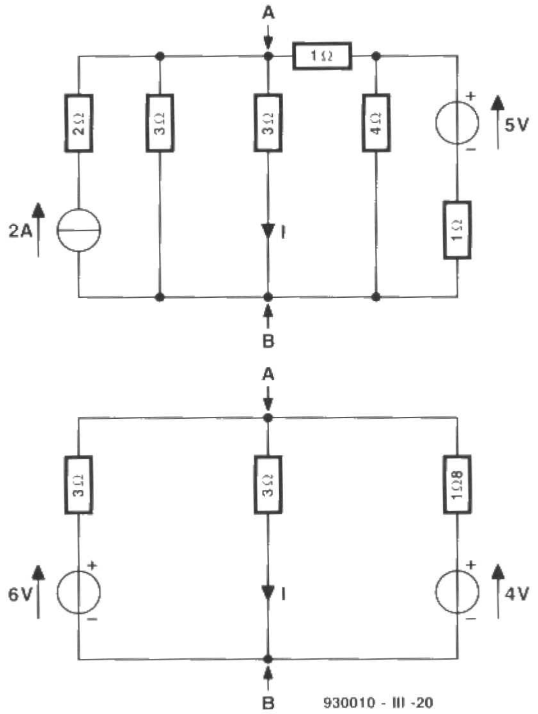


Fig. 30. A more complex example.

a

b

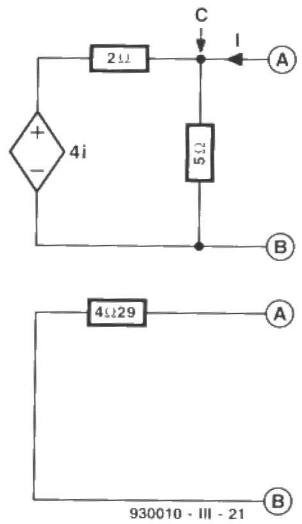


Fig. 31. Thevenin equivalent of a controlled source.

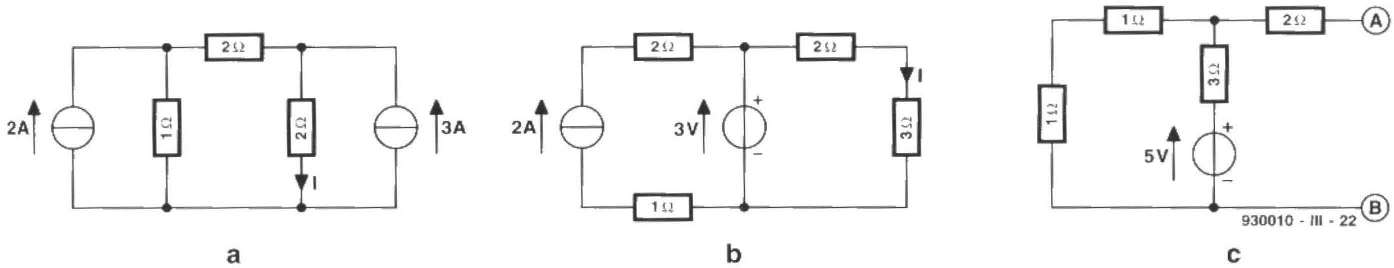


Fig. 32. Networks for Test Yourself.

A sinusoidal alternative: wave sine wave generators a goodbye

By Michael Soper, MA

In limited bandwidth circuits, any periodic or repeated waveform can more easily be made with square waves; the reason is that from a clock at twice the frequency of the upper bandwidth limit, and from the fundamental frequency, by the use of logic circuits, any harmonic can be produced.

Why move on this?

When a waveform is being constructed, for use in a synthesizer or any other use, there are many reasons for requiring

- ease of generation of harmonics;
- ease of modulation of one frequency by another.

Therefore, by using sqs(x) to denote a squared sine wave (maximum value 1; minimum value −1), for example, through a comparator, and sqc(x) to denote a square cosine wave in the same fashion, a sine wave can be written

$$\sin x=a_1\mathrm{sqs}(x)+a_3\mathrm{sqs}(3x)+a_5\mathrm{sqs}(5x),$$

where $a_j=0$ when j is even and $a_j=\pi f(j)/4j$ when j is odd [$f(j)$ is a function to determine].

Thus, to synthesize a sine wave with square waves requires the value of $1a_1$ to a_{11} and a top cut fiter as shown in the diagram, in which f is the fundamental frequency. In the diagram, the proportions are adjusted to produce a sine wave since the 'top hat' is successively rounded by stages. The final 'top cut' rounds the waveform into a sine wave.

No more troublesome sine generators

Since

$$\sin x=\pi/4[\mathrm{sqs}(x)-\mathrm{sqs}(3x)/3-\mathrm{sqs}(5x)-\ldots]$$

and

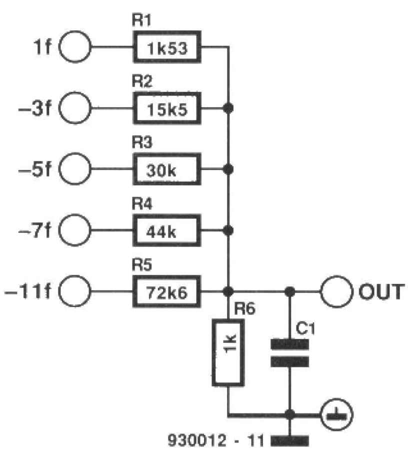
$$\cos x=\pi/4[\mathrm{sqc}(x)-\mathrm{sqc}(3x)/3-\mathrm{sqc}(5x)+\ldots],$$

where the coefficient of sqs(nx) is given by $[d(n)-3]/n$ when $n<27$ (and similarly for sqc(nx), and the forms for sin2y, sin3y, and so forth, can be obtained by substituting $x=2y$, $x=3y$, we know that by a method similar to Fourier series any repetitive waveform can be approximated by the requisite infinite series in sqs and sqc.

For example, the ramp function $f(x)$ for $-\pi<x<\pi$ can be represented by

$$(4\pi)f(x)=\mathrm{sqs}(x)-\tfrac{1}{2}\mathrm{sqs}(2x)-\tfrac{1}{4}\mathrm{sqs}(4x)-\tfrac{1}{8}\mathrm{sqs}(8x)-\ldots$$

and this approximates far more rapidly than the equivalent Fourier series. Now, is there an application in the generation of precise time-bases here? Note that a time-base generated by logic can have much greater accuracy and far greater 'repeatability' from one manufactured device to another.



The triangular wave $f(x)=\pi-2x$, where $\pi<x<\pi$, has the expansion

$$f(x)=\pi/4[\mathrm{sqc}(x)-\tfrac{1}{2}\mathrm{sqc}(2x)]-\tfrac{1}{4}\mathrm{sqc}(4x)-\tfrac{1}{8}\mathrm{sqc}(8x)+\ldots .$$

Some mathematical laws for these functions are:

1. $\mathrm{sqs}(nx)\mathrm{sqc}(nx)=\mathrm{sqs}(2nx)$;
2. $[\mathrm{sqs}(nx)]^2=1$;
3. $[\mathrm{sqc}(nx)]^2=1$;
4. $[\mathrm{sqs}(nx)][(\mathrm{sqs}(2nx)]=\mathrm{sqc}(nx)$;
5. $[\mathrm{sqs}(2nx)][\mathrm{sqc}(nx)]=\mathrm{sqs}(nx)$;
6. $\mathrm{sqs}[\mathrm{sqs}(nx)]=\mathrm{sqs}(nx)$;
7. $\mathrm{sqs}[\mathrm{sqc}(nx)]=\mathrm{sqc}(nx)$;
8. $\mathrm{sqc}[\mathrm{sqs}(nx)]=1$;
9. $\mathrm{sqc}[\mathrm{sqc}(nx)]=1$.

These results are very dissimilar to the results for sin$x$ and cos$x$ and have some remarkable features: the first point is that multiplication by sqs($2nx$) changes the phase of sqs(nx) by 90°, so that this becomes a way of converting between sqs and sqc. To perform this multiplication logically is very easy by using XOR. Let us denote XOR by v and the function of a function by o. Then

$$\mathrm{sqs}(nx)\ v\ \mathrm{sqc}(nx)=\mathrm{sqs}(2nx);$$
$$\mathrm{sqs}(nx)\ v\ \mathrm{sqc}(2nx)=\mathrm{sqc}(nx);$$
$$\mathrm{sqs}\ o\ \mathrm{sqs}=\mathrm{sqs};$$
$$\mathrm{sqc}\ o\ \mathrm{sqc}=\mathrm{sqc};$$
$$\mathrm{sqc}\ o\ \mathrm{sqs}=1;$$
$$\mathrm{sqc}\ o\ \mathrm{sqc}=1;$$
$$\mathrm{sqs}\ o\ [\mathrm{sqs}(x)\ v\ \mathrm{sqc}(x)]=\mathrm{sqs}(2nx);$$
$$\mathrm{sqs}\ o\ \mathrm{sqs}(x)\ v\ \mathrm{sqs}\ o\ \mathrm{sqc}(x)=\mathrm{sqs}(2nx).$$

FIGURING IT OUT - Part 3 (Cont'd)

from the controlled source. Solving this equation gives

$$I=0.233\ \mathrm{A},$$

from which we obtain

$$R_{Th}=U/I=1/0.233=4.29\ \Omega.$$

The Thevenin equivalent is given in Fig. 31b.

There is another useful equivalent circuit, the Norton equivalent, which we shall look at in a future issue. ■

Test yourself

1. Use the superposition method to find the value of I in Fig. 32a.
2. Replace the 2 A source in Fig. 32a with an 8 V source (positive to the top of the diagram) and recalculate I.
3. Use the superposition method to find I and U in Fig. 32b. What is the effect on I and U of replacing the 1 Ω resistor with a short circuit?
4. Exchange the voltage and current sources in Fig. 32b. What are I and U now?
5. Find the Thevenin equivalent for the circuit of Fig. 32c.
6. Remove the 3 A source from Fig. 32a, leaving an open circuit. Find the Thevenin equivalent for the remaining circuit.

Answers will be given in next month's instalment.

Note that function levels are different from logic levels.

Some similarities to sine and cosine

Since the functions given above are just squared versions of sine and cosine, the link with the traditional functions creates some similarities, for example:

$sqs(x)sqc(x)=sqs(2x)$
(like the formula for $\sin 2x$)
but
$sqc^2(x)+sqs^2(x)=2;$
$sqc^2(x)-sqs^2(x)=0;$
$2sqc^2(x)-1=1;$
$1-2sqs^2(x)=-1.$

All this is very different from the behaviour of sine and cosine.

The integral formulae for the coefficients of expansion are similar, however, and there are other useful similarities.

Reason to construct like this

1 Ease of generation of the component waveforms, which when generated are near perfect.
2. Exactly known relative phase of the waveforms.
3. Simple algebraic rules easily applied.
4. Algebra can be carried out by logic on a suitable voltage base-line.
5. Qality of the waveform is precisely determined after synthesis.
   There are many practical and natural advantages to any technical system as precisely determined as this. As has been shown, sine, ramp and triangle generators are all possible with the use of this digital method.

Mappings for the coefficients

The integrals of products of functions $f(x)$ with $sqs(x)$ and $sqc(x)$ are very easy to calculate in the knowledge of the key fact that these are both square waves of different phase. The key feature is

$$\int_0^{2\pi} sqs(nx)sqs(mx)dx = d_{mn}$$
$$= 1 \text{ iff } n = m$$
and
$$\int_0^{2\pi} sqs(nx)sqc(mx)dx = 0$$

Thus, when $f(x)$ has an sqs$m$ sqc expansion, we can work out the coefficients from $sqs(mx)$ and $sqc(mx)$ in a similar way to Fourier series. Note that in practical applications the expansion does not need to continue indefnitely and, most importantly, whenever $f(x)dx$ is known, $f(x)sqs(nx)dx$ and $f(x)sqc(nx)dx$ are easy to calculate. The formulae are

$$a_n = \tfrac{1}{2}\pi\int_{-\pi}^{\pi} sqs(nx)f(x)dx$$

$$b_n = \tfrac{1}{2}\pi\int_{-\pi}^{\pi} sqc(nx)f(x)dx$$

$$a_0 = \tfrac{1}{2}\pi\int_{-\pi}^{\pi} f(x)dx$$

Hence, the coefficients for the mapping are easily determinable. There is one point that is worth stressing: for some functions, say a periodic repetition of $\tanh(x)$ from $-\pi$ through π with period 2π, the Fourier coefficients are very difficult to determine, but the sqs and sqc coefficients are easy to calculate as long as $\int f(x)dx$ is known in general. This is naturally in very marked contrast to the difficulties of calculating the Fourier coefficients of the very many functions whose integrals become intractable when multiplied by $\sin(nx)$ or $\cos(nx)$.

Multiplying the integrand by sqs merely has the effect of changing the limits of integration of $f(x)$:

$$\int sqs(2x)f(x)dx = \int f(x)dx) +$$
$$+ \int (x)dx - \int f(x)dx - \int f(x)dx.$$

Flattening of top response

The premise of this paper is that since finite bandwidth is inevitable, generation of waveforms by synthesis with the use of sqs and sqc functions has many advantages.

Firstly, the exactly known phase of the waveform is a very great advantage, indeed, since this can be used in coherent transmission and reception techniques that are very energy saving.

Secondly, the harmonics can easily be generated by a simple counting technique based on a frequency above the top limit of the bandwidth, which is a multiple of the fundamental.

Thirdly, the calculation of the coefficients in the sqs and sqc series is simple once the indefinite integral of $f(x)$ is known.

Fourthly, a geometric series of coefficients, when this occurs, leads to the easy generation of harmonics of identical shape to the required waveform $f(\omega t)$ (by summing only the upper part of the harmonic range).

Fifthly, since the very high slop top-cut filter is relatively simple to make now (27 dB per octave is not uncommon), sine and triangular waveforms of good shape are possible.

Adopting sqs and sqc

Squaring any waveform produces a periodic squared waveform that will often, but not always, be one of $sqs(nx)$, $sqc(nx)$. However, more importantly, the precise frequency of the resulting wave is know (can be counted).

When energy-economy is considered important, the advantages of coherent transmission and reception are enormous; in fact, transmitted power can often be reduced to $^1/_{20}$–$^1/_{100}$ of the uncoherent level.

The serious question may be asked: "Why pollute the 'ether' with colossal uncoherent transmitter powers when accurate time-code transmissions are available to every transmitter in the western world, so that coherent reception is possible?"

There is a report of a short-wave transmission (from a torch battery) across the Atlantic to a coherent receiver in Britain.

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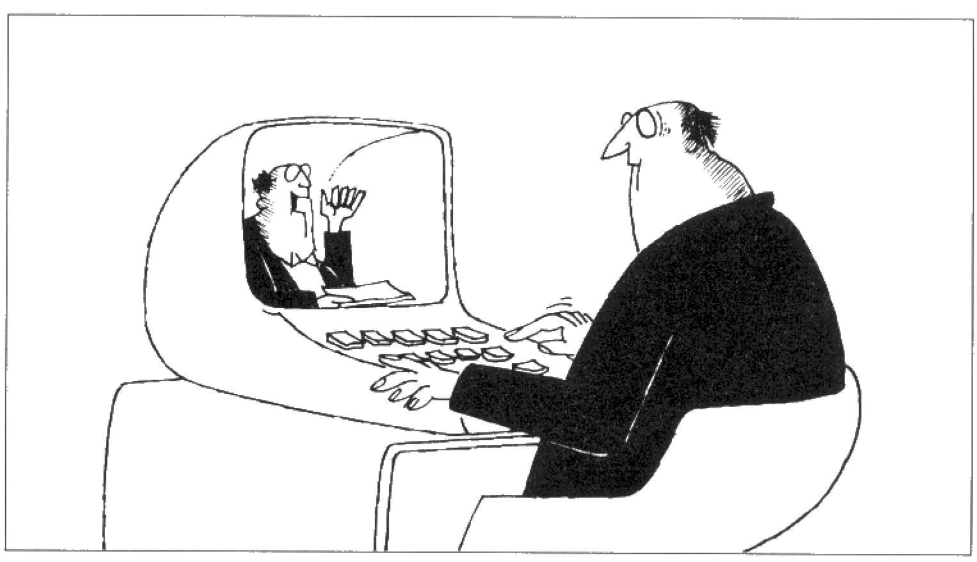


80C32 COMPUTER APPLICATION NOTES - 1

Following the eight-instalment '8051/8032 assembler course' featured in last year's issues of *Elektor Electronics*, this columm presents design ideas, programming examples and hardware experiments based on and around the popular 80C32 single-board computer. All descriptions are kept as brief as possible to ensure that a wide variety of subjects can be presented. Apart from the knowlegde you have hopefully gathered from the assembly language course, you will need the following to get going with the application notes: an 80C32 SBC (with extension board) running the EMON51 system monitor from EPROM, and a PC running the EASM51 assembler. Both programs are contained on the assembler course diskette, no. 1661. This month we kick off with a DTMF decoder and a TV test pattern generator.

DTMF decoder

Design by W. Hackländer

THE DTMF (dual-tone multi-frequency) decoder is plugged into the sockets for IC2 and IC3 on the 80C32 extension board. The ICs originally fitted at these positions are moved to the DTMF decoder board. All functions of the extension board are retained after this modification, that is, if the DTMF decoder is not selected (via software).

The input device is a commercially available hand-held DTMF tone dialler. These pocket size units are inexpensive and widely available. They are used, for instance, for remote control of answering machines. The signals emitted by the tone dialler are picked up by a microphone, amplified (IC1), decoded into a 4-bit word (IC2), and subsequently fed to the 80C32 single-board computer for further processing and displaying. An LED signals reception of a valid DTMF code. This signal may be connected to the INT0 or INT1 input of the microcontroller.

The example program written for the decoder, key.hex, reads the 4-bit DTMF word, and puts it on the LCD. Key.hex and its source code are contained on a diskette with order code 1791 (see page 70).

The heart of the circuit is formed by a Type MV8870 integrated DTMF decoder from Plessey Semiconductors (formerly a GTE Microcircuits product). The internal structure and the

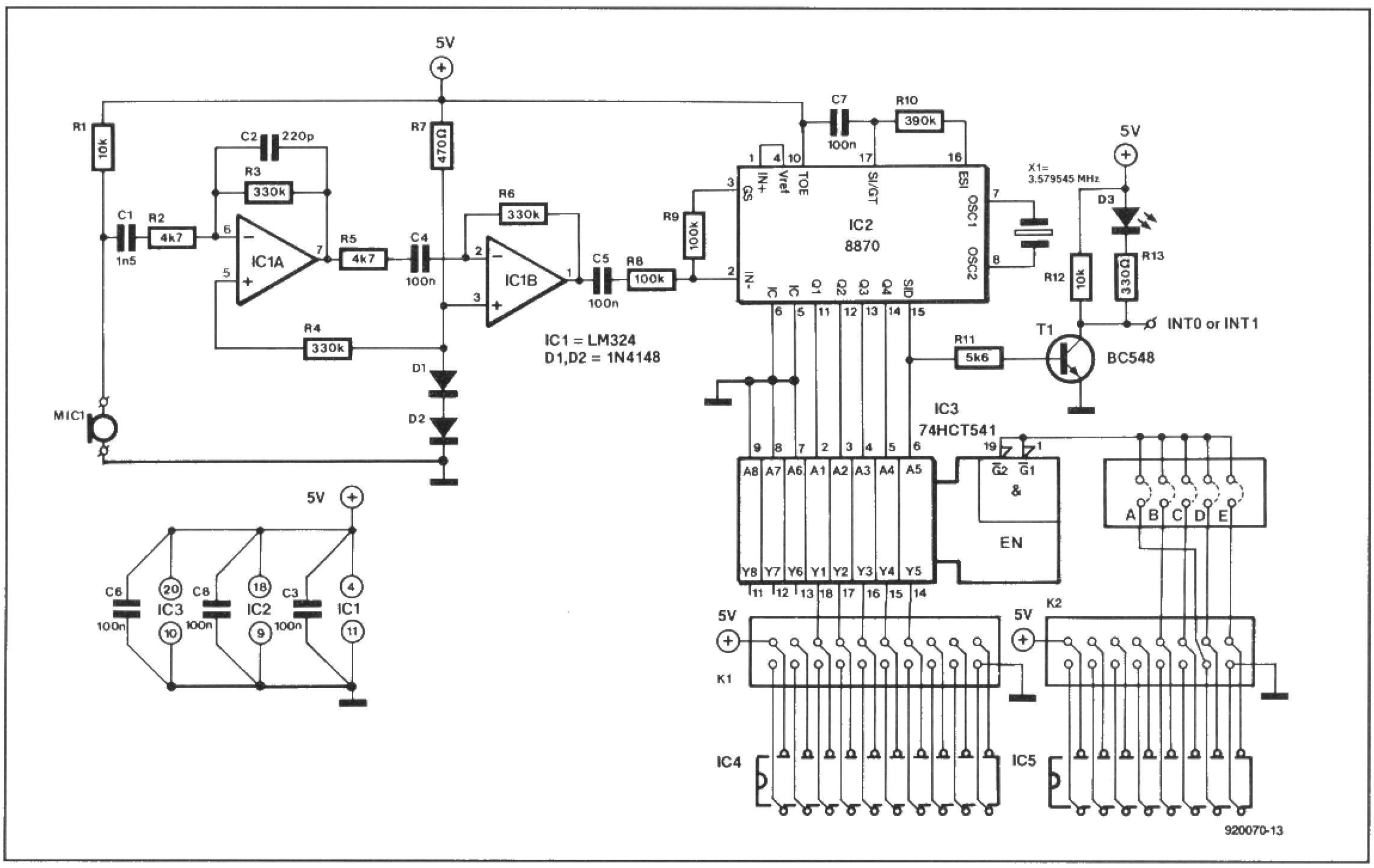


Fig. 1. Circuit diagram of the DTMF decoder. Circuits IC4 and IC5 are taken from the 80C32 extension board (where they are in positions IC3 and IC2 respectively).

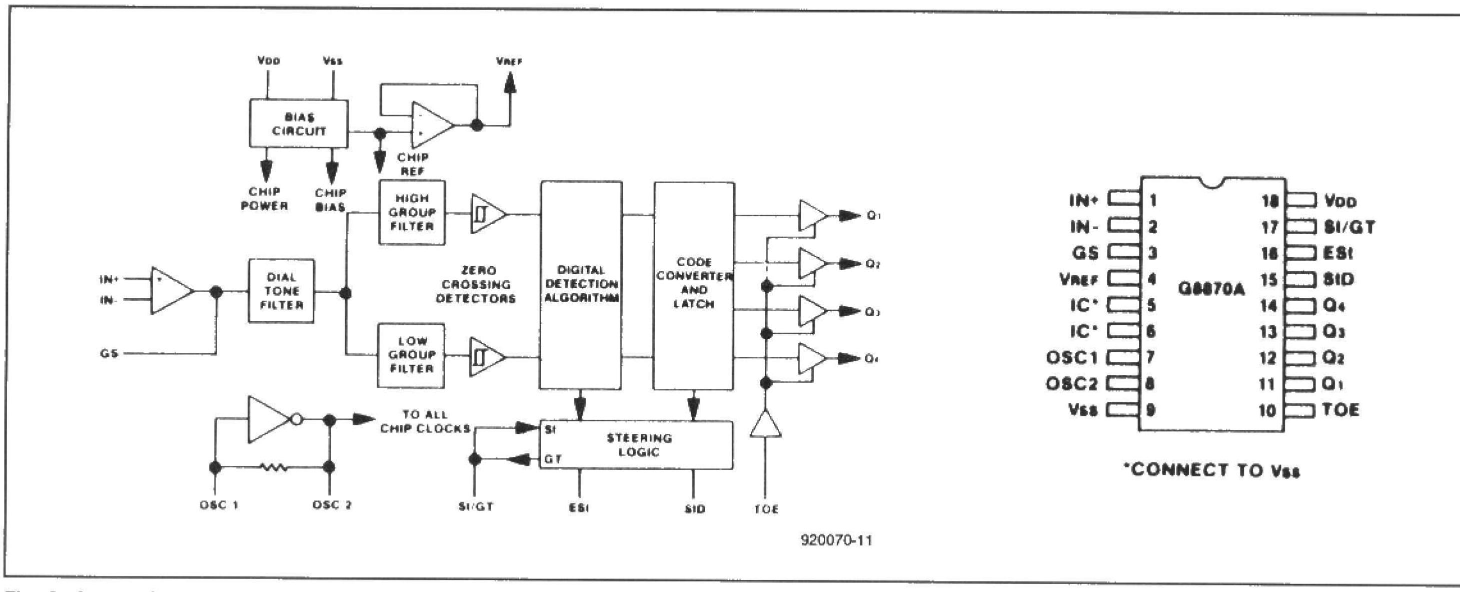


Fig. 2. Internal structure and pinout of the MV8870 DTMF decoder.

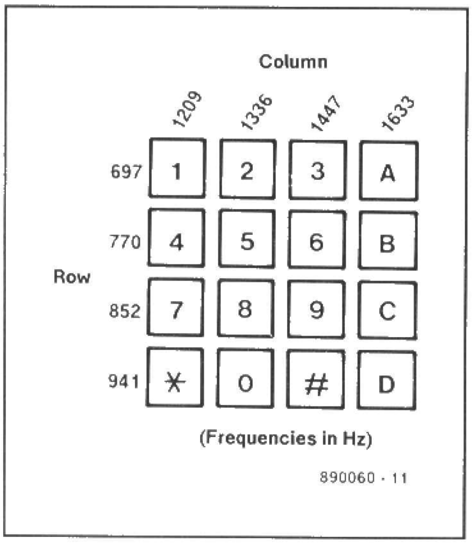


Fig. 3. Each DTMF signal consists of a low 'row' and a high 'column' tone.

pin-out of this interesting IC are given in Fig. 2. By virtue of a couple of very selective (ninth-order!) filters, the MV8870 is capable of separating the two tones that form the DTMF pair. The corresponding number, letter or sign pressed on the DTMF keypad (see Fig. 3) is computed by the block marked 'digital detection algorithm'. The filters and the digital tone recognition afford high immunity against noise, noise pulses and speech signal components.

On detecting a valid DTMF signal, the MV8870 pulls its Est (early steering flag) logic high. Next, the tone is decoded, latched, and output as a 4-bit word via pins Q1-Q4. The word remains on the outputs until a new code is recognized. The Est output, however, reverts to low again shortly after the tone disappears.

The 'steering logic' block uses Est as a control and timing signal. The Est pulse is delayed about 0.26 ms by network C7-R10. This prevents noise pulses enabling the steering control block via the ST/GT input, and causing errors. When a valid and sufficiently long DTMF signal is detected, ST/GT is made high, and the decoded word is copied into the output buffer. About 20 µs later, the STD (delayed steering flag) goes high to signal the presence of a valid DTMF word at the outputs.

Bus driver IC4 used to be IC3 on the extension board, and address decoder IC5 used to be IC2, also on the extension board. Both are removed from their sockets on the extension board, and inserted into the indicated sockets

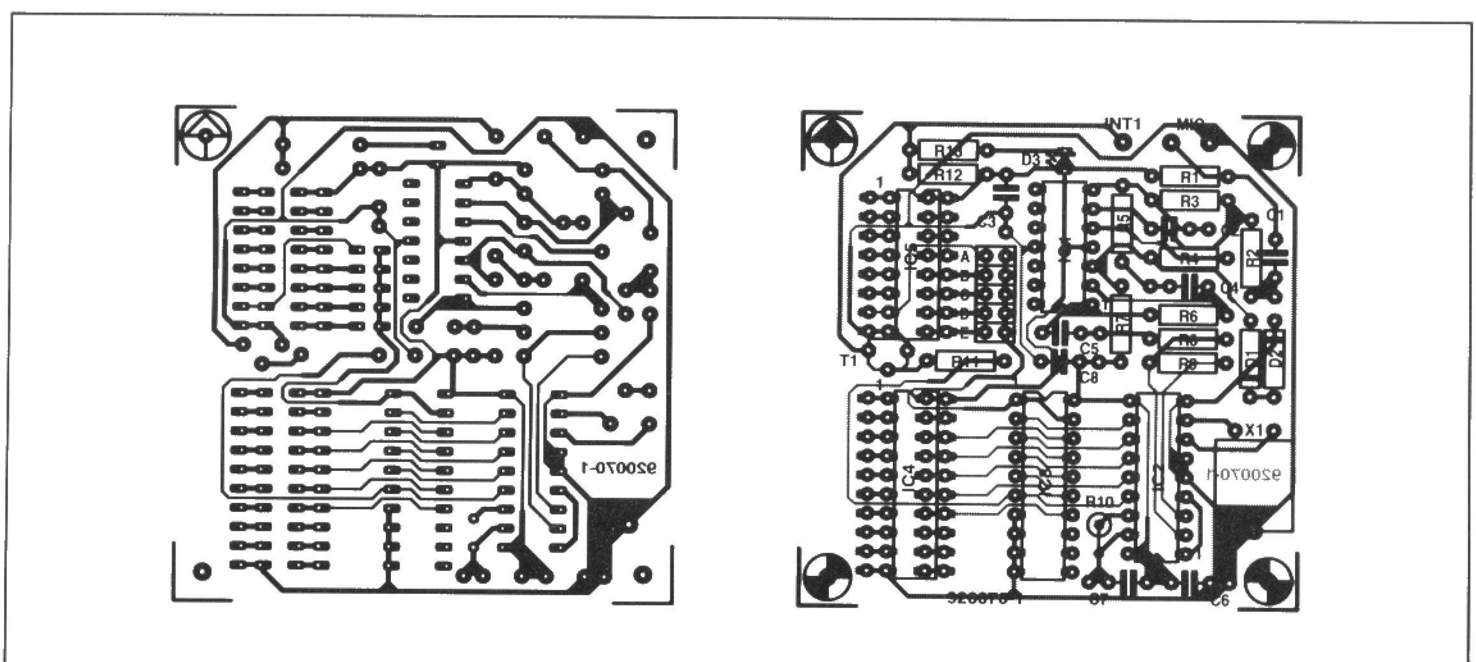


Fig. 4. The PCB for the DTMF decoder is fitted 'piggy back' on to the 80C32 extension board. Sockets with extra long pins are used in positions IC4 and IC5. These pins are inserted into sockets IC2 and IC3 on the extension board.

on the DTMF board. The empty sockets receive the pins of a 16-way and a 20-way DIP header (Fig. 5). In this way, the DTMF decoder is fitted 'piggy back' on to the extension board.

Bus driver IC4 and decoder IC3 are selected by individual three-bit addresses supplied by the computer. You determine the addresses yourself — all of them offered by Y3\ to Y7\ are available. The example program assumes that the DTMF decoder is accessed via address Y7\. In addition, fit a wire jumper at point 'A'.

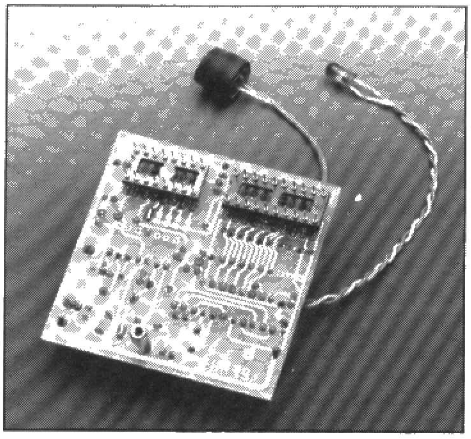


Fig. 5. Showing how the DIP headers are fitted at the solder side of the DTMF board.

COMPONENTS LIST

DTMF DECODER

Resistors:

1	10kΩ	R1
2	4kΩ7	R2;R5
3	330kΩ	R3;R4;R6
1	470Ω	R7
2	100kΩ	R8;R9
1	390kΩ	R10
1	5kΩ6	R11
1	10kΩ	R12
1	330Ω	R13

Capacitors:

1	1nF5	C1
1	220pF	C2
6	100nF	C3-C8

Semiconductors:

2	1N4148	D1;D2
1	LED, red, 5mm	D3
1	BC548	T1
1	MV8870* (Plessey) or MT8870 (Mitel)	IC2
1	74HCT541	IC3

(IC4 and IC5 are IC3 and IC2 on the extension board)

Miscellaneous:

1	Electret microphone with 2 connections	Mic1
1	3.579 MHz quartz crystal	X1
1	16-pin IC DIP pin header	
1	20-pin IC DIP pin header	

* Plessey Semiconductors Ltd., Cheney Manor, Swindon, Wilts SN2 2QW. Tel. (0793) 518000. Fax: (0793) 518411.

TV test pattern generator

Design by M. Ohsmann

To enable the 80C32 generate a TV picture, it requires a digital-to-analogue converter (DAC). This DAC must be capable of working at relatively high frequencies, typically in the MHz range. Here, the DAC is realized by an *R-2R* ladder network, which is far less expensive than a dedicated DAC IC, and does the job equally well.

The hardware part of the test pattern generator could not be simpler: apart from the *R-2R* DAC, we only need a rotary switch to select the different patterns. The block diagram, Fig. 6, indicates that the microcontroller is capable of supplying signals with an accuracy of 1 μs (12-MHz clock), provided, of course, that sufficiently fast instructions are used. The function of the port-1 bits is as follows:

- bit P1.0 is used to generate the synchronization pulse — it is low when the sync pulse is active;
- bit P1.1 is used to define the white level of the video signal, and also to generate simple black/white patterns (no grey levels);
- bits P1.4 to P1.7 drive a 4-bit DAC that supplies the staircase voltage.

The staircase signal and the sync pulse are mixed by diode D1 and pre-

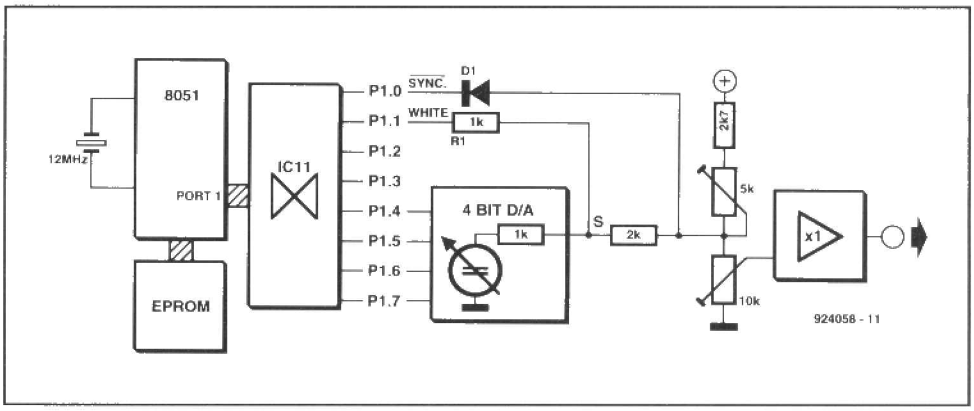


Fig.6. Block diagram of the 8032-based test pattern generator.

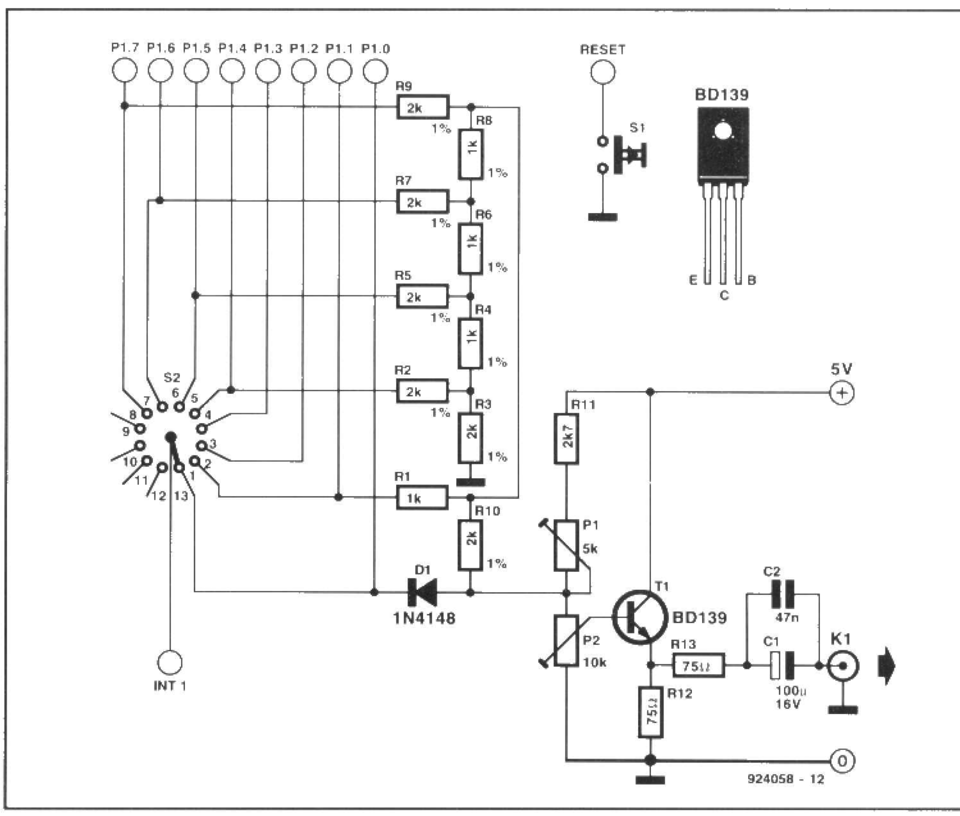


Fig. 7. A handful of inexpensive parts turn the 80C32 single-board computer into a TV test pattern generator.

```
****** LISTING of EASM51 (EVID2) ******
LINE LOC  OBJ        T          SOURCE
   1 0000                        ORG 4100H
   2 4100               P1       EQU 090H
   3 4100               ;
   4 4100 75 90 00 [2] LOOP      MOV P1,#0 ; start sync pulse
   5 4103 00       [1]           NOP
   6 4104 00       [1]           NOP
   7 4105 00       [1]           NOP
   8 4106 00       [1]           NOP
   9 4107 B2 90    [1]           CPL P1.0 ; end sync pulse
  10 4109 75 90 81 [2]           MOV P1,#081H ; middle gray
  11 410C 75 90 F1 [2]           MOV P1,#0F1H ; white
  12 410F 75 90 81 [2]           MOV P1,#081H ; gray
  13 4112 80 EC    [2]           SJMP LOOP
  14 4114                        END
********** SYMBOLTABLE (2 symbols) **********
        P1 :0090       LOOP :4100
```

Fig. 8. Example program to generate a staircase video signal.

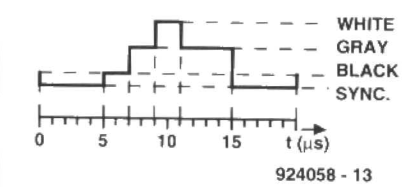


Fig. 9. This is what you get if you run the executable code produced by the listing in Fig. 8.

sets P1 and P2. The composite video signal so obtained is buffered by a single-transistor stage capable of driving a low-impedance (75-Ω) load, for instance, a monitor input.

To ensure correct signal timing, the 80C32 SBC has to run at a clock of 12 MHz. Jumper 'A' must be fitted, and choke L1 must be removed. The driver for port 1 shown in the block diagram is already available on the SBC. That means that the complete circuit of the test pattern generator can be reduced to what is shown in Fig. 7.

As an aside, the printed circuit board of the single board computer need not be fully populated for the present application. The following ICs are not required, and may be omitted: IC5 and IC6 (RAM); IC9 (address bus driver) and IC10 (data bus driver). You may also want to do without the battery backup and programming voltage circuitry. The EPROM socket, position IC7, receives the EPROM that contains the test pattern generator control program. This EPROM may be obtained ready-programmed through the Readers Services under order number 6151.

Table 1. Test pattern selection

Switch position	Test pattern
0	Staircase (grey steps)
1	Coarse chessboard
2	Fine chessboard
3	Wide vertical bars
4	Narrow vertical bars
5	Horizontal bars
6	Staircase without raster sync
7	Two vertical bars without raster sync
8	Text 'ELEKTOR'

The generator is very simple to operate: first, set the desired test pattern on switch S1 (see Table 1), and then press the RESET switch, S2. This causes the microcontroller to read the switch position (via the port lines), and run the relevant part of the control program. The levels of the synchronization and picture components in the composite video signal are set with the aid of presets P1 and P2. These adjustments are not critical, and adequate results are obtained simply by looking at the picture on the monitor.

Test patterns 6 and 7 are eminent for faultfinding in the horizontal ('line') circuity of a TV set, using an oscilloscope — the absence of a raster sync pulse keeps the scope readily triggered.

Most instructions of the 8051 (80C32) take either 12 or 24 clock cycles, which corresponds to an execution time of 1 µs or 2 µs if a 12-MHz quartz crystal is used. Figs. 8 and 9 show a small machine code program and the resultant video signal respectively. This signal has a period of 15 µs. The numbers in square brackets given in the fourth column of the listing are the instruction execution times in microseconds. The example already shows the 5-µs long synchronization pulse, and can be extended to the required TV line length of 64 µs simply by adding instructions. In this way it is possible to generate 'simple' TV picture lines. The software contained in the control program EPROM does basically the same — it repeats certain lines a number of times to create a pattern. If the instruction

MOV P1,#data

is used, changes in the grey value can occur at a resolution of only 2 µs. The control program in addition uses a number of 'set bit' and 'reset bit' instructions, which are faster (1 µs) than the MOV instruction. However, they can only set or reset one port line (P1) at a time. This is exploited as follows. Port line P1.1 is connected to resistor R1 to enable it to produce black/white transitions at a 1-µs rate, which results in better picture resolution.

Finally, port P1 has a double function in the present application. Apart from supplying the control bits to form the video signal, it is also read by the processor (after a reset) to see which line is connected to the INT1 line. On reading the relevant port bit, the processor selects the appropriate test pattern. ❑

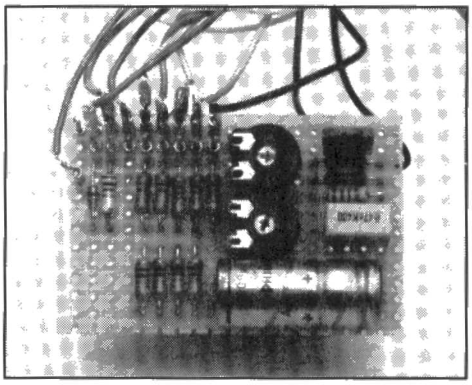
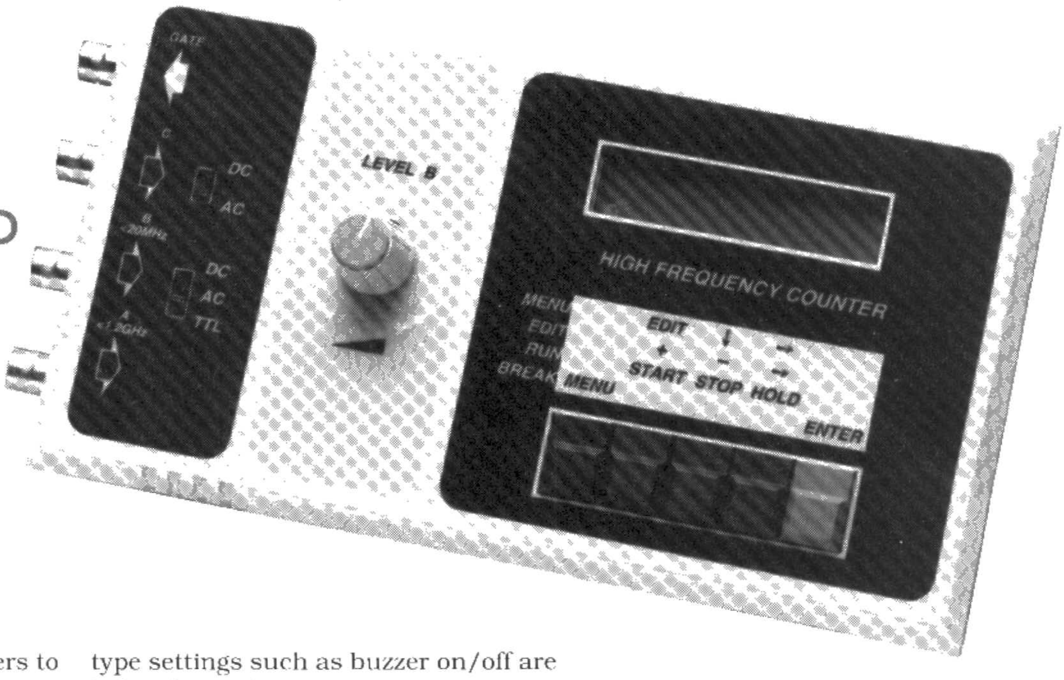


Fig. 10. The test pattern generator is easily built on a small piece of veroboard.

Next time: a variable windscreen wiper control and an RC5 (infra-red remote control) code generator.

1.2 GHz MULTIFUNCTION FREQUENCY METER



PART 4 (FINAL): THE PC LINK (CONTINUED) AND MEASUREMENT PRINCIPLES

Design by B.C. Zschocke

The PC may not transmit characters to the counter while this is busy executing the command string. Any character, in particular US, has the same effect as pressing the BREAK key on the instrument: it halts the execution of the command string, and takes the counter back to its start state (default). This may, of course, be used to break off a measurement on purpose.

By transmitting a **DC3** character, the PC prompts the counter to transmit the contents of all registers (Fig. 10i).

Control function **DC4** is used by the PC to read the current command stored in the counter (Fig. 10j). The return transmission starts with the first function contained in the command string. An ACK code indicates that the complete command has been transmitted. If the DC4 is followed immediately by ACK, the command memory is empty.

All functions contained in a command may also be executed directly, one by one (Fig. 10k). This is achieved by having the computer send the function to the counter (in connect mode). This is particularly useful when toggle-type settings such as buzzer on/off are to be changed.

A command string consists of a number of functions arranged as a sequence. On changing to command entry mode (STX), a pointer in the counter points to the first function in the command string (Fig. 10i). Any function sent to the counter is then added to the command string at the pointer position. Next, the pointer is increased by one. To check this loading process, the counter returns a copy of the stored character to the PC. If the command memory is full, the next function received is not stored, and a **GS** code is sent to the PC. Control function HT causes the function at the pointer position to be returned to the PC, and increases the pointer by one without storing the function. Control function **CAN** moves the pointer back one location, and transmits the character at the new location.

The counter returns a **NAK** code if it receives anything it can not interpret (i.e., any unknown control character or function) — see Fig. 10m.

The **RS** code allows the PC to reset the counter (Fig. 10n). This function is equivalent to switching the counter off and on again. After a reset, all register contents are undefined.

A **US** code, finally, causes the counter to revert to its start (default) state (break, Fig. 10o). At the same time, it leaves the connect or command entry mode.

Commands

A command consists of a number of individual indicators. The PC should build the command string in accordance with the structure of the menu overview shown in Fig. 8 (part 2). That is, from the top (reset) to the bottom (exit and start), with the functions preceding the parameters. As already mentioned, the relevant codes may be found in the boxes shown in Fig. 8. Table 3 lists all functions and associated codes.

An example is in order at this point to illustrate how a command string may be built. Let us assume that the following measurement is required.

Type: frequency on channel A;
Gate time: 1 s;
Start on: START key.

The string is shown analysed in Fig. 11. Also refer back to Fig. 8 to understand how the PC follows the menu structure. The two-position hexadecimal numbers are transmitted to the counter as one byte. The number of bytes per command is not fixed, since it is possible, as shown by the exam-

```
21 2F 57 3C 65 77 7F
|  |  |  |  |  |  └6th. parameter (setting: period analysis on)
|  |  |  |  |  └5th. parameter (setting: single)
|  |  |  |  └4th. parameter (start on: START key)
|  |  |  └3rd. parameter (gate time: 1 s)
|  |  └2nd. parameter (gate: gate time)
|  └1st. parameter (input: channel A)
└function: frequency measurement
```

Fig. 11. Example of a command string sent to the frequency meter by the PC.

Table 4. Function descriptions

MAIN FUNCTIONS

Frequency:	frequency measurement
1/Frequency:	frequency measurement, reciprocal indication
Revolution counter:	frequency measurement, indication in rev./min

Frequency measurement requires the following parameter functions:
- Channel...(input)
- Gate...(measurement duration)
- Gate time...(gate time, including 'gate time measured')
- Start...(start of measurement)

Pulses: count pulses
Parameter functions required:
- Channel...(input)
- Gate... (measurement duration)
- Gate time...(gate time, only with 'Gate preset')
- Start...(start of measurement)

Time: time measurement
Parameter functions required:
- Gate on key pressed or start/stop key

Timer: timer device
Parameter functions required:
- User def. period (loads time to be set)
- Start...(start of measurement)
- Pulse...(pulse on output)

Pulse generator
Parameter functions required:
- Preset period duration or preset pulse/pause
- User def. period or
- User def. duration and user def. pause
- Start...(Start)
- End...(Stop)
- User def. pulse (with 'end on number of periods')

Zero counter
Zero counter/manual counter
Parameter functions required:
- Start value... (start value)
- Channel C L-H or channel C H-L (input)
- Active... or pulse on EQ0 or count pulse (output)
- Start on...(Start)

PARAMETER FUNCTIONS

Signal input

Channel A signal input:	channel A
Channel B signal input:	channel B
Channel C signal input:	channel C
Channel C H signal input:	channel C, active high-phase
Channel C L signal input:	channel C, active low-phase
Channel C H-L signal input:	channel C, active H-to-L edge
Channel C L-H signal input:	channel C, active L-to-H edge

Gate time (Gate)

Gate time 0.1 s	Load gate time register with 0.1 sec.
Gate time 1 s	Load gate time register with 1 sec.
Gate time 10 s	Load gate time register with 10 sec.
Gate time 1 min	Load gate time register with 1 min.
Gate time user def.	Gate time to equal contents of gate time register
Gate time measured	Gate time measured
Gate time rising	Gate time increases with each measurement

Preset times for pulse generator

Preset period duration	Period duration is preset (in period duration register)
Preset pulse/pause	Pulse duration and pulse pause are preset in corresponding registers
User def. period	Load period register
User def. duration	Load pulse duration register
User def. pause	Load pulse/pause register
User def. pulse	Load register with number of pulses

Start value

Start value nought	Preset start value register with 0 and load.
Start value user def.	Load start value register

Measurement duration (Gate)

Gate preset	Gate time defined by gate time register
Gate channel C high	Gate time defined by high pulse on channel C
Gate channel C low	Gate time defined by low pulse on channel C
Gate channel C H-L	Gate time defined by H-to-L transition on channel C
Gate channel C L-H	Gate time defined by L-to-H transition on channel C
Gate key pressed	Gate time as long as START key pressed
Gate start/stop key	Gate time starts on START key, and ends on STOP key

Start on (Start)

Start immediately	Start measurement/pulse generator immediately
Start on signal	Start measurement on detection of signal
Start on signal A	Start measurement/pulse generator on detection of signal on channel A
Start on signal B	Start measurement/pulse generator on detection of signal on channel B
Start on channel C H-L	H-to-L signal transition on channel C starts measurement/pulse generator
Start on channel C L-H	L-to-H signal transition on channel C starts measurement/pulse generator
Start on START key	Start key starts measurement/pulse generator

End on (Stop)

End on no. of periods	End after predetermined number of periods
End on channel C H-L	End on H-to-L transition on channel C
End on channel C L-H	End on L-to-H transition on channel C
End on STOP key	End when STOP key pressed

Output (Timer)

Pulse on start/end	Output pulse on start and end
Pulse on start	Output pulse on start
Pulse on end	Output pulse on end
Pulse f. start to end	Output active from start to end

Output (Manual/Zero counter)

Active when NE0	Output active as long as count $\neq 0$
Active when EQ0	Output active as long as count $= 0$
Pulse on EQ0	Output pulse when counter reaches state 0
Count pulse	One output pulse per count pulse

The following functions may be executed directly or as parameter functions:

Measurement order	Change from 'continuous' to 'single' and the other way around.
Continuous	Continuous series of measurements
Single	Single measurement
Buzzer	Toggle buzzer on/off
Beep on	Switch on buzzer
Beep off	Switch off buzzer
Intermediate value	Switch between 'interm. result displayed' and 'interm. result not displayed' (toggle).
With interm. result	Switch on intermediate result display function
Without interm. result	Switch off intermediate result display function
Period analysis	Change between 'Period analysis on' and 'Period analysis off' (toggle)
Period analysis on	as is
Period analysis off	as is
Pulse polarity	Change pulse polarity
Pulse polarity pos.	Pulse polarity is positive
Pulse polarity neg.	Pulse polarity is negative
Inactive level	Change pulse inactive level
Inactive level low	Pulse inactive level is low (0 V)
Inactive level high	Pulse inactive level is high (+5 V)
Main Break	Do BREAK (counter changes to basic mode/settings)
Reset	Do RESET
Run command	Execute command
Buzzer	Beep!

REGISTER DESIGNATIONS (all values unsigned)

Direct registers (R1,..., Rn)

R1	Frequency measurement: real measurement (gate-) time in µs (after measurement)
	Pulse counter: real measurement (gate-) time in µs (after measurement)
	Time measurement: measured time in µs
	Pulse generator: preset pulse duration in µs
	Timer: remaining time in µs
R2	Pulse generator: period duration (if preset) in µs
	Timer: preset time in µs
	Manual/Zero counter: start value
R3	Frequency measurement: number of pulses counted, (prescaler ignored). Measured frequency computed from: R3/R1
	Pulse counter: number of pulses counted (prescaler ignored)
	Pulse generator: preset pulse pause duration in µs
	Manual/zero counter: counter state
R4	Reserved
R5	Pulse generator: number of pulses to be generated

Indirect registers (I1,...,In)

I1	Reserved
I2	Frequency measurement: result (as on LCD readout) without exponent
I3	Reserved
I4	Reserved
I5	Frequency measurement: preset time in µs

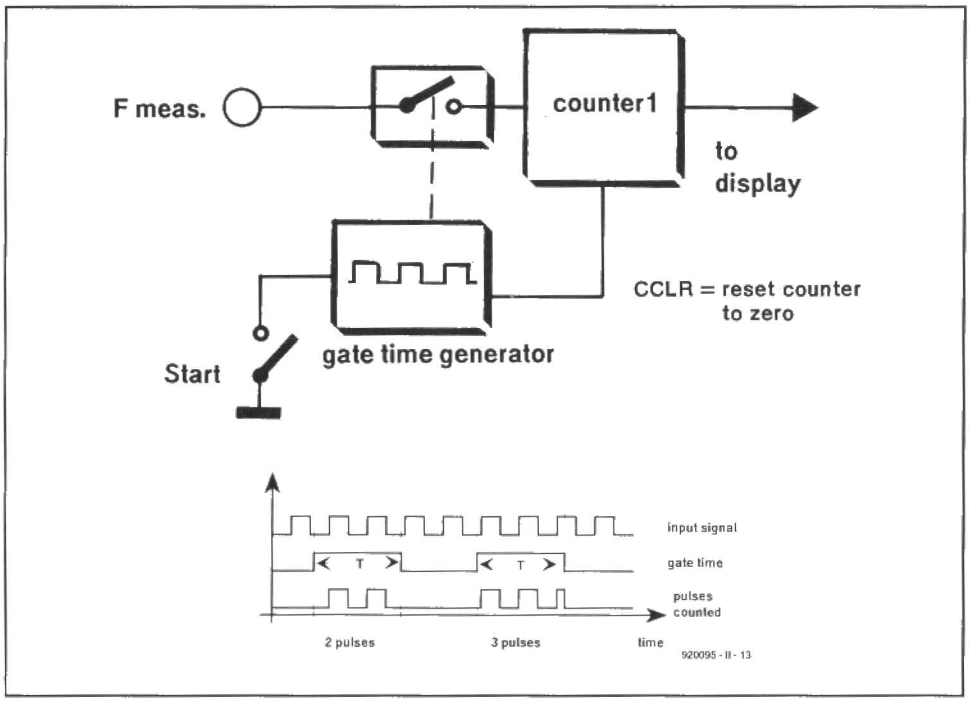


Fig. 12. The 'classic' digital frequency meter counts the periods of the input signal for a predefined time.

ple, that more than one parameter is required to complete the **settings**.

The counter executes the command from the right to the left, i.e., the **settings** before the **functions**.

You may not use parameters that are shown without a box code (Fig. 8), or that are marked with an asterisk in Table 3. Since the counter does not run a 'plausibility check' on received command strings, the user must make sure that these consist of meaningful parameters. This is not difficult to ensure by virtue of a useful trick that may be used during program development: simply use manual control to give the instrument the desired settings, and then read out the string using the DC4 command.

The frequency measurement principle

In an earlier instalment of this article it was stated that a separate instalment was to be devoted to the frequency measurement principle used by the instrument. The division of the complete article into instalments having taken a slightly different form than originally planned, we have decided to include this information in the present (final) instalment.

The usual way of measuring frequencies is to count the number of pulses that occur within a predefined gate time. Although this is not the most accurate method, it is by far the simplest. The number crunching power of a microcontroller or microprocessor, however, allows us to devise much more advanced measurement methods, of which practical applications may be found in Refs. 1 and 2. Although the same measurement principle is used for the frequency meter function of the 1.2-GHz multifunction frequency meter, this instrument makes even more use of combined software and hardware possibilities offered by the microcontroller. Also, there are now two counters instead of one counter and a programmable divider (of which the setting is determined beforehand by running a 'sample' measurement). The second counter replaces the programmable divider (in digital design, dividers and counters are often considered identical components). The nice thing about this new setup is that the sample measurement is no longer required, which results in a shorter measurement time. To understand how this works, it may be useful to recap the operation of the pulse counting principle used in 'classic' frequency meters.

The classic approach

To refresh your memory, Fig. 12 shows the architecture of the classic, pulse counting, frequency meter. A clock circuit supplies a gate signal that serves to connect the input signal to the counter for an accurately determined time, T. The number of pulses N counted during the gate time T thus gives the input signal frequency ($F=N/T$).

The accuracy of the measurement is determined by two factors: first, the accuracy of the gate time, and, secondly, the number of pulses counted. The latter factor is responsible for the relatively low accuracy at low frequencies. As illustrated by the timing diagram in Fig. 12, there may be an error of one in the number of pulses counted. As shown, it all depends on how the gate time, T, coincides with the periods of the input signal. The resulting absolute error, Δ_{abs}, is calculated from

$$\Delta_{abs} = 1 \text{ (pulse)} / T \text{ (s)} \quad \text{[Hz]}$$

Consequently, the measured frequency may have an error of 1 Hz at a gate time of 1 s, and 10 Hz at a gate time of 0.1 s. This error becomes smaller as the frequency increases, when the main cause of errors is increasingly on account of gate signal deviations. The table below shows the effect of the counting error at a gate time of 0.1 s:

f	Δ_{abs}	Δ_{rel}
1 MHz	10 Hz	0.001%
1 kHz	10 Hz	1%
10 Hz	10 Hz	100%

Frequencies lower than 10 Hz are not given simply because they can not be measured at a gate time of 0.1 s. Inevitably, lower frequencies require longer measurement times, which brings us to another disadvantage of the classic frequency measurement principle: measuring low frequencies accurately takes a lot of time.

Ratio-based measurements

A measurement principle that is eminently suited to microprocessor implementation is shown in Fig. 13. The basic principle is very simple. A certain time is reserved to measure the periods of the input signal and those of the reference frequency. Dividing the two gives the ratio of the input frequency and the reference frequency. Multiplying this ratio with the reference frequency then yields the frequency of the input signal.

If we say "a certain time", this has to be taken literally, since the gate time is really only an auxiliary signal in this setup. The input signal frequency is calculated exclusively on the basis of the counter states N :

$$f = f_{ref} \, (N_1 / N_2).$$

Bear in mind, however, that the measurement has to run for at least one period of the input signal.

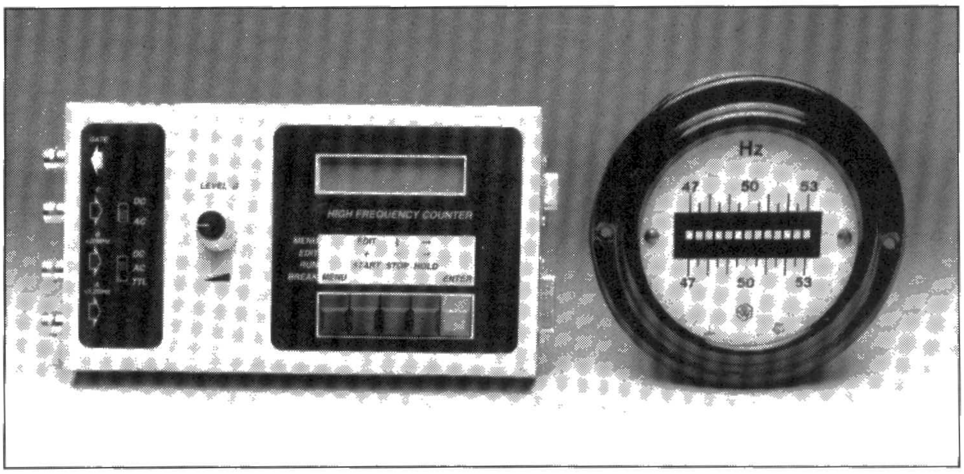
The fact that the gate time is an independent parameter, opens up the possibility to use the computer for 'fine tuning' of the result, or, in other words, make the measurement a little more accurate. This is necessary anyway because both counters make an error of one pulse if the gate time were

Table 3. Function/code overview.

Main functions	
* First Number	020H
* Measurement function	020H
Frequency	021H
1/Frequency	022H
Rev counter	023H
Pulse counter	024H
* Reserved	025H
* "	
* Reserved	028H
Time	029H
Timer	02AH
Pulse generator	02BH
Zero counter	02CH
Manual counter	02DH
Parameter functions	
* Input	02EH
Channel A	02FH
Channel B	030H
Channel C	031H
Channel C high	032H
Channel C low	033H
Channel C high-to-low	034H
Channel C low-to-high	035H
* Ratio	036H
Period duration	037H
Pause period	038H
Pause duration	039H
* Gate time	03AH
Gate time 0.1 sec	03BH
Gate time 1 sec	03CH
Gate time 10 sec	03DH
Gate time 1 min	03EH
Gate time user def.	03FH
Gate time measured	040H
Gate time rising	046H
Preset period duration	047H
Preset pulse/pause	048H
User def. period	049H
User def. duration	04AH
User def. pause	04BH
User def. pulse	04CH
* Number of periods	04DH
* Reserved	04EH
* "	
* Reserved	052H
* Start value	053H
Start value nought	054H
Start value user def.	055H
* Gate	056H
Gate preset	057H
Gate Channel C high	058H
Gate Channel C low	059H
Gate Channel C high-to-low	05AH
Gate Channel C low-to-high	05BH
Gate key pressed	05CH
Gate START-STOP	05DH
* Start of measurement	05EH
Start immediately	05FH
Start on signal	060H
Start on signal A	061H
Start on signal B	062H
Start on channel C high-to-low	063H
Start on channel C low-to-high	064H
Start on START key	065H
* End	066H
End on no. of periods	067H
End on channel C high-to-low	068H
End on channel C low-to-high	069H
End on STOP key	06AH
* Output	06BH
Pulse on start/end	06CH
Pulse on start	06DH
Pulse on end	06EH
Pulse from start to end	06FH
Active when NE0	070H
Active when EQ0	071H
Pulse on EQ0	072H
Count pulse	073H
Counter setting functions	
* Miscellaneous	074H
Measurement order	075H
Continuous	076H
Single	077H
Buzzer	078H
Beep on	079H
Beep off	07AH
Intermediate value	07BH
With interm. result	07CH
Without interm. result	07DH
Period analysis	07EH
Period analysis on	07FH
Period analysis off	080H
Pulse polarity	081H
Pulse polarity pos.	083H
Pulse polarity neg.	084H
Inactive level	085H
Inactive level low	086H
Inactive level high	087H
* Reserved	088H
Remaining functions	
* Input	089H
* Reserved	08AH
* "	
* Reserved	092H
Main Break	093H
Reset	094H
* System	095H
* Reserved	096H
* Reserved	097H
Run command	098H
Buzzer	099H
* Reserved	09AH

* = not significant for PC control

indeed arbitrary. This potential problem is solved by having the computer adjust the gate time such that counter 1 processes a whole number of periods. This rules out errors in the number of pulses counted by counter 1. The timing diagram in Fig. 13 shows what happens. After the gate time T has elapsed, the system keeps counting for a time Δt, so as to include the input signal period that has just started. Unfortunately, the above 'trick' can not be applied to counter 2. That is nothing to worry about, however, provided the counter is fed with a great many pulses. If this is so, the error introduced by the single missing pulse is considerably reduced, as already explained in the section on the classic frequency meter. The number of pulses to be counted by counter 2 depends on the reference frequency (f_{ref}) and the gate time. The reference frequency being fixed (500 kHz in the case of the multifunction frequency meter), it will be obvious that we must maintain a reasonably long gate time (the shortest gate time that can be set on the instrument, 100 μs, is just about acceptable).



Although an error of one pulse is inherent in the operation of counter 2, there is still a means of increasing the accuracy of the measurement. To begin with, we have the computer provide a fixed logic level (for instance, 0) at the input of the counter when the gate time starts. This is achieved with a software-controlled inverter. In this way, we are certain that the first (already running) period of the reference signal is included in the count as long as possible. Also, we can have the computer check the logic level of f_{ref} at the end of the gate time. In fact, this produces an error that is smaller than one pulse. All in all, we can safely assume an error of one pulse for the error calculation. The relative error made by counter 2 is $1/N_2$. To ensure the smallest possible relative error, N_2 must be as large as possible. This can be achieved by making f_{ref} and/or the gate time as large as possible.

Returning to the measured frequency calculation, $f = f_{ref}\,(N_1/N_2)$, you may spot another source of errors: the reference frequency. The relative error in this frequency is determined by the quartz crystal used to generate the reference clock. The total measurement error thus becomes:

$$\Delta_{rel} = \Delta f_{ref} + 1/N_2.$$

To calculate the error, it is easier to write $f_{ref}\,(T + \Delta t)$ instead of N2, because f_{ref} is known, T is set on the instrument, and Δt is negligible at relatively high frequencies, and easily calculated at low frequencies on the basis of the measurement result. In addition, the more extensive notation indicates clearly that the relative accuracy of the measurement depends exclusively on (1) the reference frequency, (2) its stability, and (3) the time reserved for the measurement, instead of on the measured frequency.

So, what does it all do in the case of the instrument described? Assuming a measurement time of 0.1 s and a reference frequency accuracy of, for instance, 100 ppm (0.01%), the relative error is as small as

$$\Delta_{rel} = 0.01\% + 100\%/(500\text{ kHz} \times 0.1\text{ s}) = 0.012\%$$

or 120 ppm. Obviously, the relative error is even smaller if the stability of the reference frequency is better, and the measurement time longer.

The functions indicated in Fig. 13 are not easily found back in the circuit diagram of the instrument (Fig. 2 in part 1). In fact, only the gate signal (on connector K5) and a piece of counter 1 are obvious, the rest is implemented by the hardware contained in the microcontroller. ■

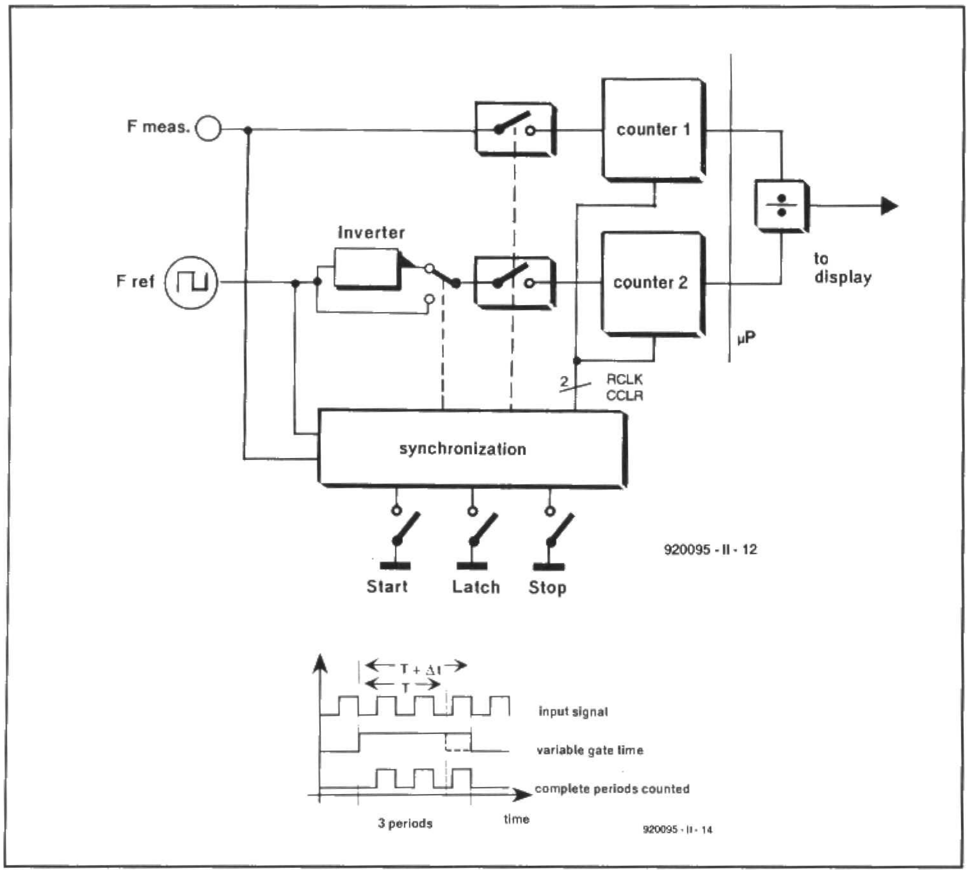


Fig. 13. By virtue of the dual-counter approach, a computer-based frequency meter achieves greater accuracy than a 'classic' design (compare Fig. 12).

References:

1. 'Microprocessor-controlled frequency meter'. *Elektor Electronics* January 1985.

2. 'Multifunction measurement card for PCs' *Elektor Electronics* January and February 1991.

READERS' CORNER

LETTERS

8051/8032 Assembler Course

Dear Editor—In the 8051/8032 Assembler Course, the software No. 1661 contains the EMON51.HEX file. Could you please tell me what type of hex file it is, since I cannot convert it into a bin file as used by my EPROM programmer.

S.K. Pang, Harlow, Essex.

It is important to note that the EMON51.HEX file is in a format designed by the author. It ***cannot*** *be used as a straight hex file to send to an EPROM programmer. It is an intermediate file format, generated by the assembler and used to produce an Intel file and a binary file (simply look at what MAKE1 does). To make sure you have the correct EMON 51.HEX file, we recommend that you assemble EMON51.A51 using EASM51.*

If your diskette no. 1661 does not contain a subdirectory called \FILECONV, you have an early release. Please return it for a free update to

J. Buiting
Central Design Department
Elektor Electronics
P.O. Box 75
6190 AB Beek
The Netherlands
Telephone +31 46 389444
Fax +31 46 370 161.

If you have an Intel compatible programmer, file conversion may not be necessary at all. The simplest thing to do is to send the file "intel.hex" (in the \fileconv subdirectory on your disk) straight to your EPROM programmer.

Alternatively, assemble EMON51.A51, and run the batch file "MAKE1". This will produce two files: "intel.hex" and "binary.bin". When comparing the generated "intel.hex" file with the one in \fileconv, you may find that the two are different by one byte. This byte represents a date number at location 02A1. It is either 39 or 35, which can give EPROM checksums of 1357 or 135B (the date reads 15 August or 19 August). Both are acceptable.

If you have little experience in working with EPROM programmers and their input file formats, we strongly recommend that you obtain a ready-programmed EPROM from us. The order code is 6061, and the EPROM comes with the course disk. The EMON51 monitor works at 4800 baud.

[Technical Editor]

Transputer Module

Dear Editor—I was very impressed with your 8032 computer card (May 1991), which gave me the idea of doing something a little different. My transputer module (TRAM) has: T225 16-bit processor, 10 MIPS, 20 MHz internal clock; 64K static memory, single wait state; size 2 single-sided PCB and only 5 wire links (cf. Inmos Design 35 MHz, 64K zero wait, size 1 4-layer PCB, £270). This may be suitable for an *Elektor Electronics* project. The main problem is that to program such a module, you need: IBM PC, Atari, Sun, or VME computer system; TRAM motherboard for the PC (£650); 2 Mbyte 32-bit TRAM with subsystem (£1000); language program, e.g., Occam, C or Fortran (£500–£1000).

There are lots of other bits for this transputer system that I could design extremely fast. I can think of another two projects that are a lot simpler and could be prototyped on matrix board in a couple of hours.

10 MIPS is quite a powerful processor. Transputers, unlike conventional processors, can be plugged together to increase the overall processing power. Make ten, and it would cost £600 for 100 MIPS. Unfortunately, you have to program in Occam to use these processors to their full potential.

There are a few corners that could be cut to make a transputer system less expensive. E.g., Inmos's backplane specification could be down-graded to just a 5 MHz clock and a link chip that sits PC I/O space. This chip costs less than £10. Then, instead of having

EVENTS

IEE AND IEEIE PROGRAMME

1 Mar—Learning from the Piper Alpha disaster.
8 Mar—Total quality ISO9000.
9 Mar—16th Edition update.
10 Mar—Periodic inspection and testing of electrical installations.
10 Mar—Writing good technical reports.
11 Mar—Portable appliance testing.
17 Mar—Electromagnetic compatibility for project managers.
18 Mar—Language learning for engineers and technicians.
22 Mar—Data communications and the transport industry.
23 Mar—Electricity at work regulations.
30 Mar–2 Apr—Eighth International Conference on Antennas and Propagation (ICAP 93)

Further information on these, and many other, events may be obtained from the IEE, Savoy Place, London WC2R 0BL; Telephone 071 240 1871, or from the IEEIE, Savoy Hill House, Savoy Hill, London WC2R 0BS; Telephone 071 836 3357.

DIGITAL SIGNAL PROCESSING

ERA Technology's conference and exhibition on digital signal processing, 'DSP – The Enabling Technology for Communications' will take place at the RAI Congress Centrum in Amsterdam on 9–10 March.

Details from Janine Wilson, ERA Technology, Cleeve Road, Leatherhead KT22 7SA, England. Phone (0372) 374 151

NAB '93

The 1993 National Association of Broadcasters' exhibition and symposium will be held in Las Vegas on 19–23 April. This international event offers the full spectrum of new—and future—broadcasting, post-production, HDTV and multimedia technology and trends. From radio and television managements and programming to engineering...from advertising to sales and marketing...from multimedia to post-production.

Further information from NAB, 1771 N Street N.W., Washington, D.C. 20036-2891. Telephone +1 202 775 4972; fax +1 202 775 2146.

ASIA TELECOM 93

The Asia Telecom 93 Exhibition and Forum will be held in Singapore from 17 to 22 May 1993 under the theme 'Telecommunity: the next era of growth'. Hosted by Singapore Telecom and the Telecommunication Authority of Singapore, Asia Telecom 93 is organized by the International Telecommunication Union

Further information from Asia Telecom 93, ITU, Place des Nations, CH-1211, Geneva 20, Switzerland. Phone +41 22 730 5811.

NETWORLD EUROPE '93

Networld Europe '93, the pan-European showcase and discussion forum for networking and connectivity will take place in Frankfurt on 25–27 May, 1993.

Details from Saryl Leifeld at Blenheim International, Düsseldorf, Germany. Phone +49 211 901 9186.

SPRING 93 COMPUTER SHOPPER SHOW

Blenheim PEL's Spring 1993 Computer Shopper Show will be held in the National Hall, Olympia, on 27–30 May 1993.

NETWORKS '93

Networks '93, the exhibition for the information technology (IT) industry, will take place in Hall 5 of the National Exhibition Centre, Birmingham from 29 June to 1 July 1993.

Details on these two events from Blenheim PEL, 630 Chiswick High Road, London W4 5BG, England. Telephone 081 742 2828.

a network of reconfigurable TRAMs, it could be just a fixed array shape. TRAM sockets are merely 16-pin DIL IC sockets with about 3.3 in between the rows of pins.

The modification would require some software changes, but the source code is provided with the Transputer Development System.

Total system costs could be reduced still further by designing your own 32-bit 2 Mbyte TRAM. These transputers are really clever: they even have their own memory refresh circuitry and programmable positions of memory timing. The ones sold by Inmos are very good, but expensive. A little degrading of the specification would quickly reduce the price. More TRAMs could be added later to improve the processing speed.

R. Cooke, Wigan.

Thank you for this interesting proposal. However, owing to its very limited appeal, we do not think that it is a suitable project for further development and publication by us. Nevertheless, since we feel that it may appeal to some readers, it is reproduced here in full. Interested readers are invited to write to Mr Cooke via our Editorial Offices in Dorchester. [Technical Editor]

COMPONENT RATINGS

In resistor and capacitor values, decimal points and large numbers of zeros are avoided wherever possible. Small and large values are usually abbreviated as follows:

p (pico-) $= 10^{-12}$
n (nano-) $= 10^{-9}$
µ (micro-) $= 10^{-6}$
m (milli-) $= 10^{-3}$
k (kilo-) $= 10^{3}$
M (mega-) $= 10^{6}$
G (giga-) $= 10^{9}$

Note that nano-farad (nF) is the international way of writing 1000 pF or 0.001 µF.
Resistors are $^1/_3$ watt, 5% metal film types unless otherwise specified.

The direct working voltage of capacitors (other than electrolytic or tantalum types) is assumed to be ≥60 V. As a rule of thumb, a safe value is about 2× direct supply voltage.

Direct test voltages are measured with a 20 kΩ/V meter unless otherwise specified.

Mains (power line) voltages are not listed in the articles. It is assumed that our readers know what voltage is standard in their part of the world.

Readers in countries that use 60 Hz supplies, should note that our circuits are usually designed for 50 Hz. This will not normally cause problems, although if the mains frequency is used for synchronization, some modification may be required.

The international letter symbol '*U*' is used for voltage instead of the ambiguous '*V*'. The letter V is reserved for 'volts'.

The size of a metric bolt or screw is defined by the letter M followed by a number corresponding to the overall diameter of the thread in mm, the × sign and the length of the bolt or screw, also in mm. For instance, an M4×6 bolt has a thread diameter of 4 mm and a length of 6 mm. The overall diameter of the thread in the BA sizes is: 0 BA = 6.12 mm; 2 BA = 4.78 mm; 4 BA = 3.68 mm; 6 BA = 2.85 mm; 8 BA = 2.25 mm.

SWITCHBOARD

Switchboard allows all PRIVATE READERS of *Elektor Electronics* one FREE advertisement of up to 108 characters, including spaces, commas, numerals, etc., per month.

Write the advertisement, which MUST relate to electronics, in the coupon on this page; it MUST INCLUDE a private telephone number or name and address; post office boxes are NOT acceptable.

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FOR SALE. IBM PC, 24-line I/O card, £ 30. Speech, music synthesizer and others connected to it. Phone Pei at Manchester 061 272 8279.

HELP. Can anyone give me information on the Zycor 4000 computer? Please ring Tom on 021 707 3906.

FOR SALE. 2×28-way PCB edge connectors, right-angled; 5 for £1. Amstrad CPC mouse interface: 2 for £1. Telephone Richard on 0525 210 126.

FOR SALE. BBC Master with Z80 2nd proc. colour monitor, twin disc drives, games, Wordstar. Offers. Please phone 0883 342 306.

WANTED. Circuit diagram or other information for RCA Receiver AR88. Please contact Mr Hardy G3KND, QTHR, or phone 0252 26824.

WANTED. January–April, incl.1987 and January–November, incl. 1991 copies of *Elektor Electronics* in good condition. Please write to Jarkko Laukkanen, Helenivksenk, J6 A17, SF-05860, Hyvinkáá, Finland.

WANTED. Manual or copy for Nordmende UW342U wobbulator or loan to copy. Reasonable price willingly paid. Write to Mr W. Mansell, 48 Bowling Green Road, Thatcham RG13 3DA, England.

FOR SALE. Small qty RF valves: QQV06-40A, £15; QQV03-10A, £5; JAN829B, £15. All new and boxed. Phone 0249 445 069.

HELP. Where can I buy the Spoerle slider switches for the Elektor RS232 tester (July 1992). Call Jim on (0624) 824 380.

HELP! Circuit or manual for Telequipment S32A oscilloscope. Please write to Iain Ferguson, 88 Sheepcote Lane, London SW11 5BP.

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ELEKTOR ELECTRONICS 03/93

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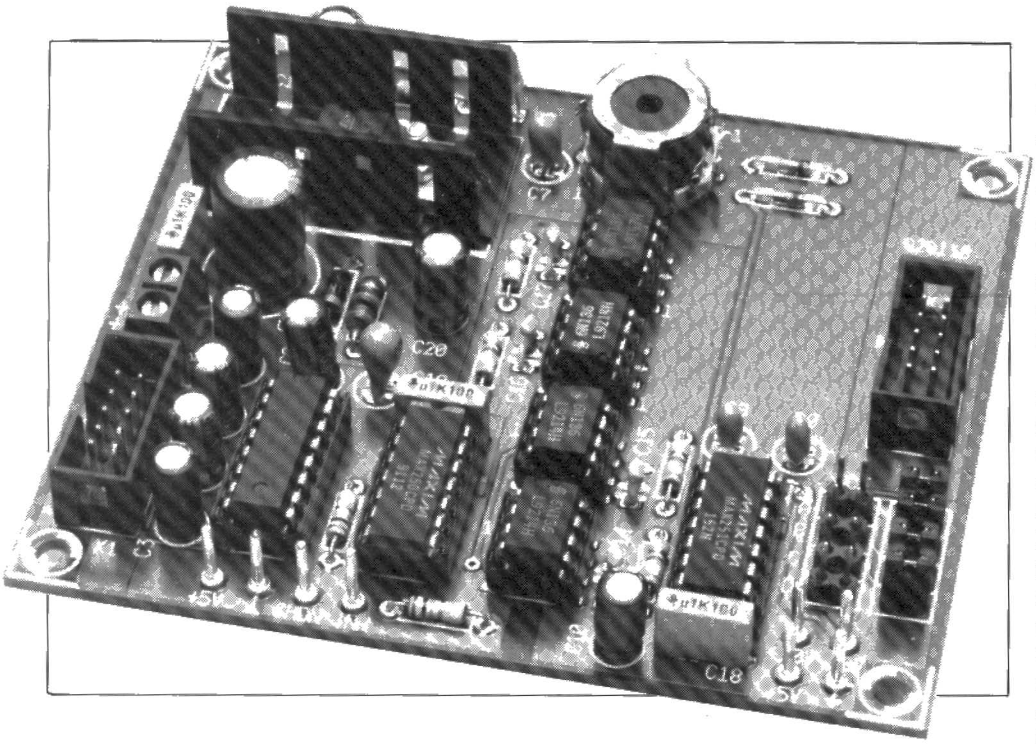
ELECTRICALLY ISOLATED RS232 INTERFACE



The cable is by far the most widely used type of connection where electrical signals are to be exchanged between two pieces of equipment or larger systems. An inherent disadvantage of a cable is, however, that it brings the electrical potentials of the two systems in contact with each other, which can have many undesirable electrical as well as electromagnetic effects. The latter, in particular, are not to be waved aside, witness the recently enforced EMC (electromagnetic compatibility) regulations, aimed at reducing the levels of interference emanating from today's electrical equipment. With the aid of an electrically isolated interface it is still possible to exchange signals without a 'copper-to-copper' connection. Such an interface is described here: it is suitable for insertion into an RS232 link as used, for instance, in an automated, remote-logging data acquisition system, where it should help to reduce measurement errors caused by common-mode interference, which is always around.

Design by J. Ruiters

Specifications

Supply voltage:	8-15 V (mains adaptor)
Current consumption:	approx. 75 mA
Max. data speed:	19,200 baud
Bandwidth:	50 kHz
Handshaking:	software (Xon/Xoff) or hardware (RTS/CTS)
Transmission:	jumper setting DTE/DCE or DTE/DTE

AN electrically isolated interface consists of a transmitter and a receiver. A marked characteristic of such a device is that the communication between the transmitter section and the receiver section does not make use of electrical conduction. Two different 'carriers' are available to convey information 'without wires': light (in optocouplers), or magnetic coupling (in isolation transformers).

Isolating devices are often applied to meet safety regulations. As such, they are aimed primarily at the well-being of people, but also at preventing damage to equipment. Typical examples include electrical isolation between low-voltage and high-voltage sections in medical equipment, and the linking of data communication systems with the aid of optocouplers or fibre optics. Optical transmission media offer better protection against static discharges, and also prevent electrical fault conditions propagating from one piece of equipment to another. The latter characteristic also enables appliances with different ground potentials to be interconnected without problems.

Apart from ensuring safety there is a second, less known but certainly not less important, reason for an electrical isolation device being desirable in certain cases. The practical cases we are referring to are fairly complex because the isolating device is used mainly to reduce the effect of a common-mode noise source. Since the origins of this type of interference may not be known to many of you, they are discussed in some detail below.

Common-mode noise

Before turning to the suppression of common-mode noise, it is useful to investigate the source of this interference and its possible, negative, effects. For a clear insight into the problem (common-mode signals), we make use of theories developed in the field of electromagnetic compatibility (EMC).

The first point to note when looking at the origin of common-mode currents (or voltages), is that we always have to take the environment of an electrical circuit into consideration. This means that the circuit model shown in Fig. 1 does not exist in practice. None the less, the model is widely used for calculations, simply because its electrical behaviour is determined

mainly by the components used, which makes the environment a negligible factor. Since common-mode noise sources are a result of a non-ideal circuit environment, the model in Fig. 2 is extended with components C_p and L_a.

In Fig. 2, the stray capacitance C_p and self-inductance L_a (which may result, for instance, from a safety earth line) represent the coupling between the (desired) circuit and its environment. The stray capacitance also indicates that the circuit and its environment are also coupled in the absence of a properly conducting connection between the circuit and its environment.

The generator (U_G, R_G) and the load (R_L) found in the diagram represent various equipment and/or circuit sections, ranging from outdoor lamps to video cameras. In most cases, the 'environment' that matters is the safety earth line, but it can also be a metal table top, or a central heating radiator — it all depends on which type of coupling is dominant in a particular frequency range.

The way in which a large common-mode noise source can come into existence is illustrated in Fig. 3, where the voltage across R_L is measured with a remote-controlled data acquisition system. If the system is within the coverage area of a transmitter (which is nearly always the case), a voltage (interference) is induced in the shaded loop shown in Fig. 4. The way in which this interference affects the operation of the measurement system is indicated in Fig. 5: a common-mode source U_{cm} inserts itself in the loop. As a result of U_{cm}, currents start to flow in the system reference (indicated by the ground symbol) and in the 'hot' wires. The term 'common mode' is explained by the fact that the interference currents in the signal wires and the reference conductors may, at any time, have the same direction. More serious than the name might indicate at first glance, however, is the fact that U_{cm} and Z_1 (the local impedance of the system reference) give rise to a noise voltage across resistor R_L (see Fig. 6). Further, it will be found that Z_2, too, contributes to the measured voltage (U_m) being determined to some extent by U_{cm} (impedance Z_2 is the local impedance of the system reference between the load resistance and the input of the measurement system).

In the above case, it is readily seen that improving the accuracy of the measurement simply entails reducing the effect of the common-mode source. A number of alternatives are available to achieve this. The most evident concept in this case is to minimize junction impedances Z_1 and Z_2. Depending on the common-mode frequency and the length of the connecting wires (with a self-inductance of 1 nH/mm), this measure will, however, yield adequate results in a limited number of cases only. If the loop area can not be reduced, it is possible to reduce the common-mode current (starting from the low-frequency approach) by inserting a high impedance in the common-mode loop. As illustrated by Fig. 7, this can be achieved by breaking the loop,

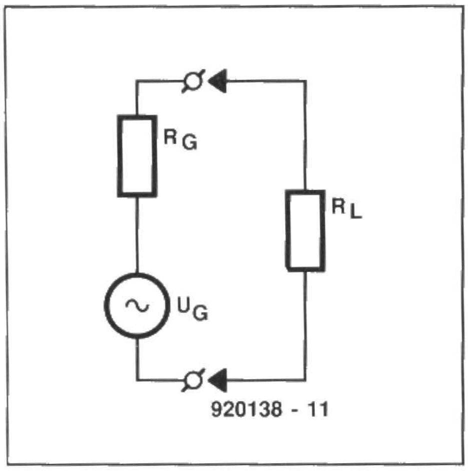


Fig. 1. Basic electrical model of a generator connected to a load.

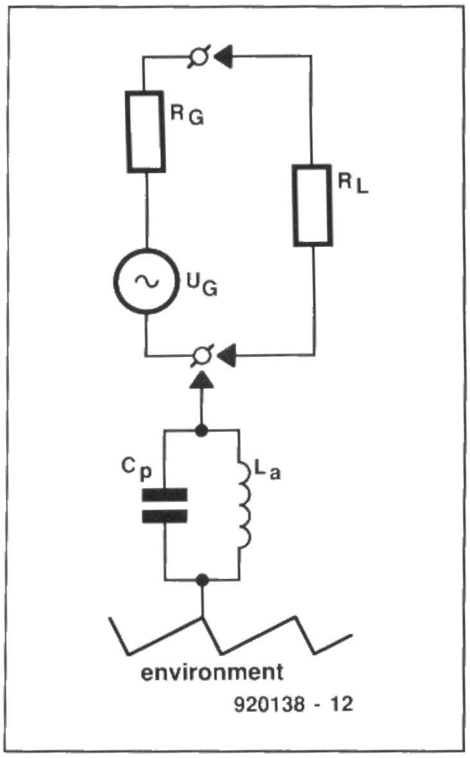


Fig. 2. Every circuit is linked to its environment, in one way way or another.

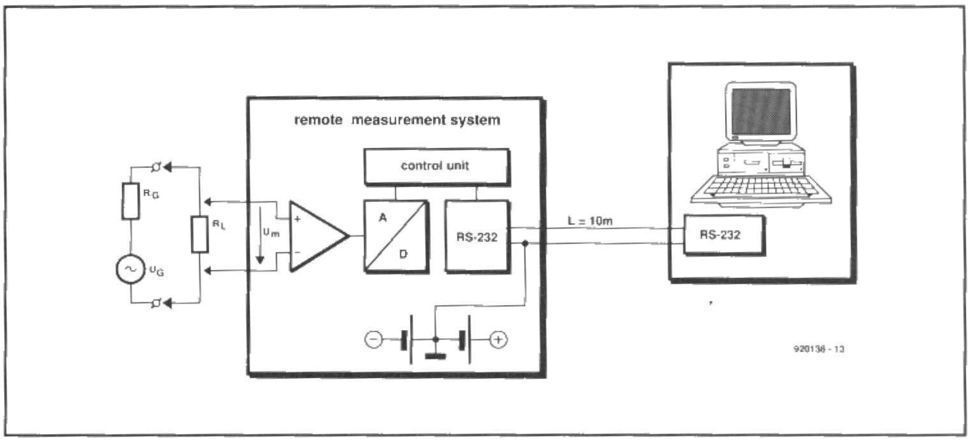


Fig. 3. Measurement setup consisting of a generator (with load), a measurement circuit and a PC for remote results processing.

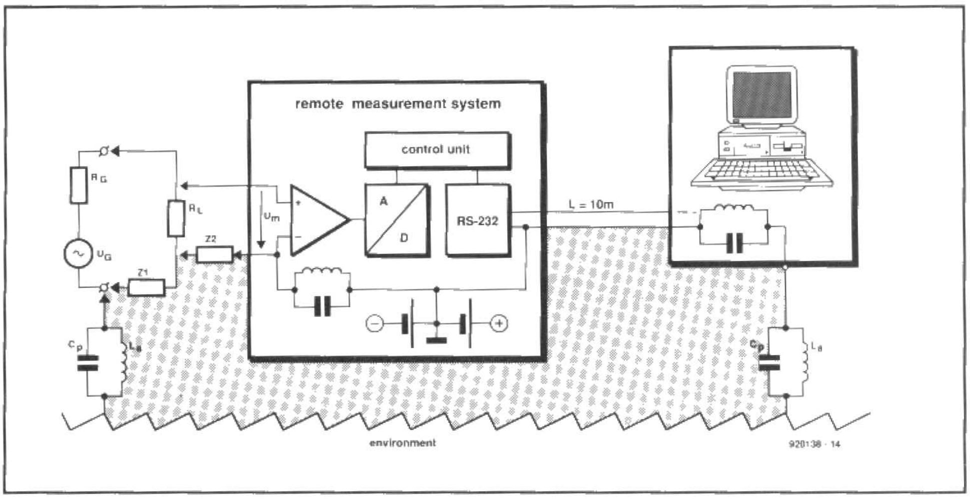


Fig. 4. In the measurement setup, too, all components are linked to one another via the environment.

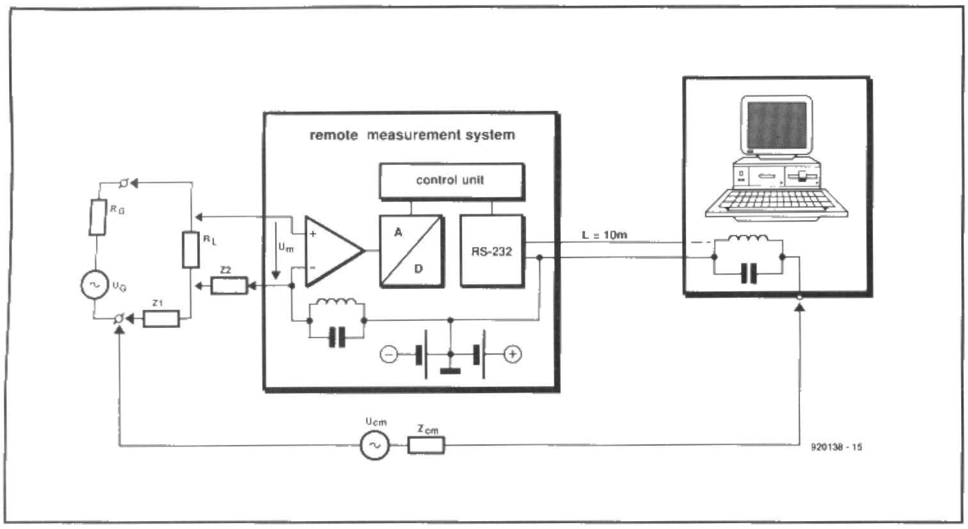


Fig. 5. The long connection in the measurement system introduces a common-mode signal that causes errors.

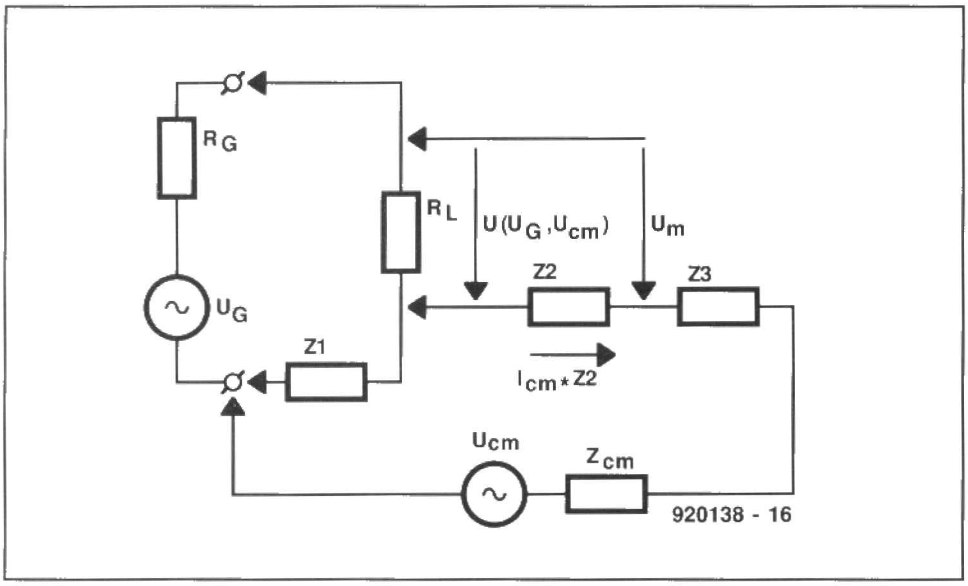


Fig. 6. Noise voltages are caused by common-mode currents that flow through local impedances *Z1* and *Z2*.

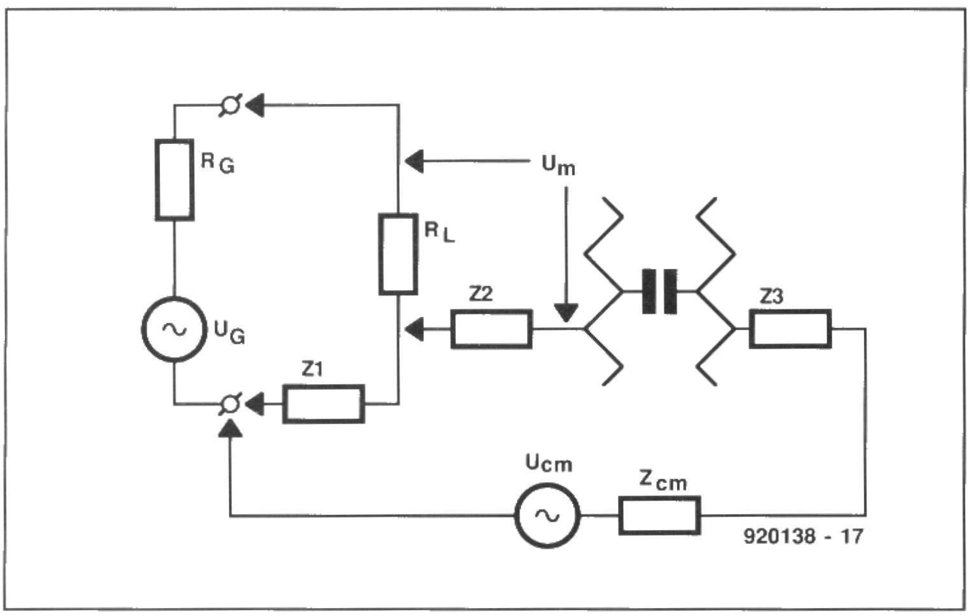


Fig. 7. The negative effects of the common-mode signal source may be eliminated by ensuring the best possible isolation. Here, a high impedance is inserted into the loop.

which is essentially the same as inserting a series capacitance having the smallest possible value.

The circuit

The best solution to the above problems is to apply an isolating device. Common-mode signals may be suppressed effectively by using a magnetic or optical coupling device. Figure 8 shows the circuit diagram of the optical isolator. RS232-compatible signals are applied to the input ('DTE side'; data terminal equipment) of the circuit, while the output ('DCE side'; data communication equipment) supplies RS232 signals. In this circuit, three integrated circuits from Maxim Inc. play an important role in the processing of the signals. Their functions are supported by a number of optocouplers and a 'tailor-made' pot core transformer. The transformer is used to ensure complete isolation of the power supplies used at the receiver and the transmitter side. Likewise, the optocouplers enable data to be exchanged between the two blocks without the need for an electrical connection.

Circuit IC1 is used to convert RS232 compatible signals into TTL levels, and vice versa. An important advantage of the two-way conversion is that the quality of the signals and their edges can be guaranteed because they can actually be improved (by 're-shaping') significantly.

A powerful feature of the MAX232 is that it is capable of generating the required positive and negative supply voltages (±10 V), using a single input supply voltage (5 V) and four external capacitors.

The MAX250 (IC2) serves to make the TTL signals in the circuit suitable for driving the optocouplers. The current source outputs of this IC are perfect for driving the LEDs in the optocoupler. Furthermore, the IC has two FET outputs, which are useful for driving transformer Tr1. The transformer is a pot-core type that provides electrical isolation of the DCE/DTE (modem) side of the circuit. The transformer is driven in push-pull mode at a frequency of 150 kHz. The MOSFETs contained in IC2 alternately take one end of the primary winding of Tr1 to ground. As required by the push-pull arrangement, the duty factor of the switching signal equals 0.5. Each primary inductor (1-2; 2-3) consists of 6 turns, while the secondary inductor (4-6) has twice as many. This gives a transformer ratio of 1:2. Since the transformer is fed by a 5-V rectangular wave, the secondary winding supplies a rectangular-shaped alternating voltage of 20 V_{pp}. By using a rectifier (D1 and an internal diode between pin 1

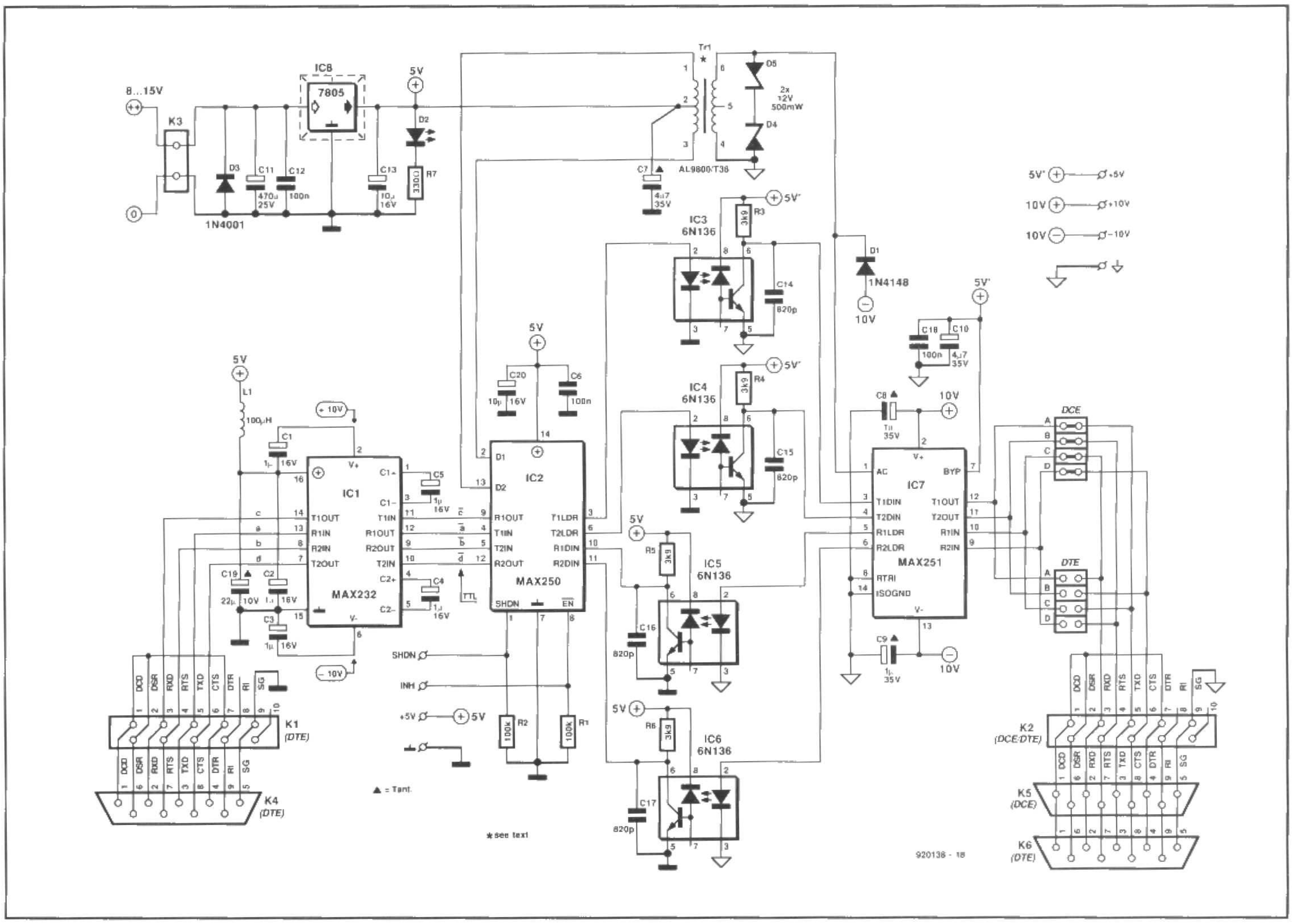
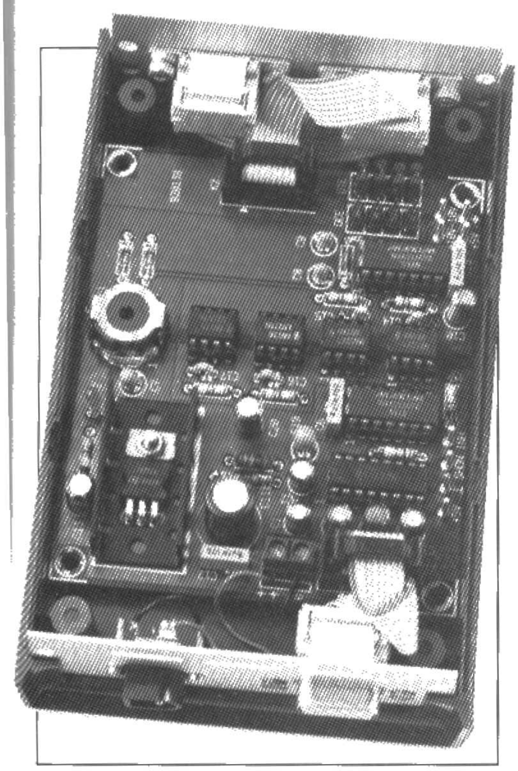


Fig. 8. By virtue of the use of three integrated circuits, the RS232 isolator remains a relatively simple and compact circuit.

and 2 of IC7), this voltage is easily converted into the correct supply levels. Four jumpers enable you to select between a DTE-DCE connection (computer-to-modem) or a DTE-DTE connection (zero-modem).



Capacitors C14 to C17 have a clearly defined function: they serve to guarantee a high common-mode suppression, also at relatively high frequencies.

Another important argument for the use of an electrical isolation device is to eliminate an earth or reference loop by breaking it. We must hasten to add, however, that such a loop can only be broken electrically. Unfortunately, since there will always remain some stray capacitance between electrically isolated sections, a loop can never be eliminated over the full frequency spectrum. This means that the (desired) suppression of common-mode signals will depend on their frequency. In the case of the optocoupler, the effective signal transfer is via light. Apart from this optical coupling, there are stray (parasitical) types of coupling that weaken the suppression of common-mode signals as their frequency increases. Figure 9 shows a model of an optocoupler with stray capacitances C_{pb} (between the primary side and the base of the transistor), and C_{ps} (between the primary and the secondary side). As a result of stray capacitances, common-mode currents are converted into a differential-mode (DM) voltage. This is a very undesirable effect since the DM voltage ends up in series with the desired signal voltage, which could cause the (digital) MAX251 to produce wrong output levels.

Considering the suppression of the common-mode signals, it is important to reduce the bandwidth of the optocoupler (Type 6N136; approx. 250 kHz) to a value that is high enough for the transmission of RS232 signals. Here, capacitors C14 to C17 limit the bandwidth of the RS232 isolator to about

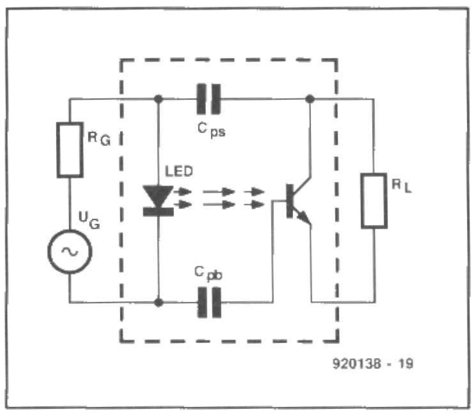


Fig. 9. Each and every optocoupler contains small series capacitances that can give rise to interference.

50 kHz, which is ample for transmission rates of up to 19,200 baud (which equals a signal frequency of 9,600 Hz).

Construction

As shown in Fig. 10, a double-sided PCB is used, with reference (ground) areas in strategic places. Three of these areas are found at the DTE (computer) side: one for the switching part of the push-pull converter, one for the optocouplers, and one for the charge pump contained in the MAX232. Since the MOSFETs switch the transformer currents to pin 7 of IC2, the ground areas make contact around this pin only. This results in a 'clean' reference for the optocouplers and the RS232 signals. The modem side also has three reference areas: one to carry the return currents produced by the transformer secondary, one to reduce the cross-talk between the optocouplers, and one (below K2) to keep the common-mode currents produced by the 'modem' device at the edge of the PCB.

With the possible exception of winding the special transformer, the construction of the circuit should not cause problems. The pot core assembly used to make the transformer is a Siemens type. The parts that form the assembly are shown in Fig. 11. The primary and secondary inductors must be wound by hand, which requires great care and accuracy. The important thing about the transformer is that the primary inductors are simultaneously wound on the holder. If this is not done, the resulting unequal inductors cause unequal fields that will neutralize each other only partly. This may lead to core saturation. Simultaneous winding can be achieved by using a 40-cm long piece of 'litz' (multi-strand) wire, which is folded double. Use this double wire to wind six turns in one compartment of the holder. To prevent the inductors coming loose, fix the conductors in place using a narrow strip of self-adhesive tape. The inductors made in this way are shown schematically in Fig. 12. Please note that the common end, marked B-C, is a result of folding the wire, and can not be used as a centre tap on the transformer. To make sure that the currents flow in the right direction (remember, the fields are to compensate one another), the common end is cut open, and the centre tap is created as shown in Fig. 12: after cutting the fold, simply connect one of the wire ends to one of the free ends.

The secondary inductor is much easier to make. It consists of 12 turns of litz wire wound into the empty compartment of the holder. Secure the winding using a bit of self-adhesive tape. Finally, assemble the trans-

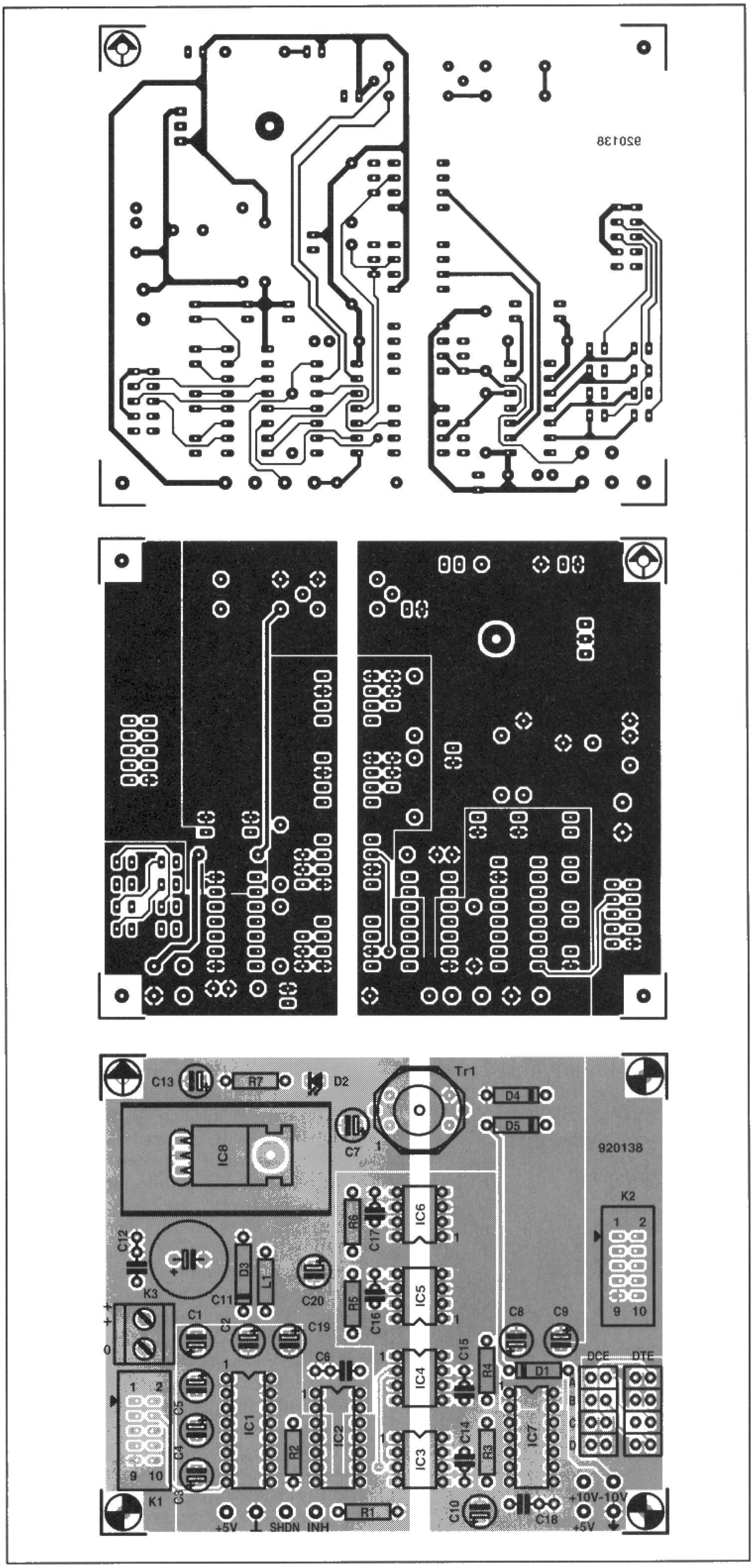


Fig. 10. Printed circuit board design for the RS232 isolator.

COMPONENTS LIST

Resistors:

2	100kΩ	R1;R2
4	3kΩ9	R3-R6
1	330Ω	R7

Capacitors:

5	1µF 16V radial	C1-C5
3	100nF	C6;C12;C18
1	4µF7 35V tantalum	C7
2	1µF 35V tantalum	C8;C9
1	4µF7 35V radial	C10
1	470µF 25V radial	C11
1	10µF 16V radial	C13
4	820pF	C14-C17
1	22µF 10V tantalum	C19
1	10µF 16V radial	C20

Semiconductors:

1	1N4148	D1
1	LED 3mm red	D2
1	1N4001	D3
2	BZX79C12 (12V/500mW zener diode)	D4;D5
1	MAX232**	IC1
1	MAX250**	IC2
4	6N136	IC3-IC6
1	MAX251**	IC7
1	7805	IC8

Miscellaneous:

2	10-way box header	K1;K2
1	2-way PCB mount terminal block, pitch 5mm	K3
2	9-way IDC style sub-D socket	K4;K6
1	9-way IDC style sub-D plug	K5
1	100µH choke	L1
	Approx. 60cm litz-wound wire dia 0.3mm	
1	Heat-sink ICK35SA (Fischer***)	
1	Case, 145×40×90 mm. E.g., Retex**** Elbox RE1	
1	Printed circuit board 920138 (see page 70)	
1	Pot core transformer assembly* Siemens type AL9800/T38	Tr1

* Parts required for pot core assembly:
- one core (B65541-W-Y38); material T38; AL value 9800nH; size 14×8mm; without air gap.
- one holder with two compartments (B65542-B-T2).
- two isolation disks (B65542-A5000).
- one connection plate with yoke (B65545-B10).

Primary inductors: 6 turns. Secondary inductor: 12 turns.

* ElectroValue, 28 St Jude's Road, Englefield Green, Egham, Surrey TW20 0HB. Telephone: (0784) 433603. Fax: (0784) 435216.
** Maxim UK distributors: 2001 Electronic Components (0438) 742001; HB Electronics Ltd. (0204) 25544.
*** Dau Components (0243) 553031.
**** Boss Industrial Mouldings Ltd. (0638) 716101.

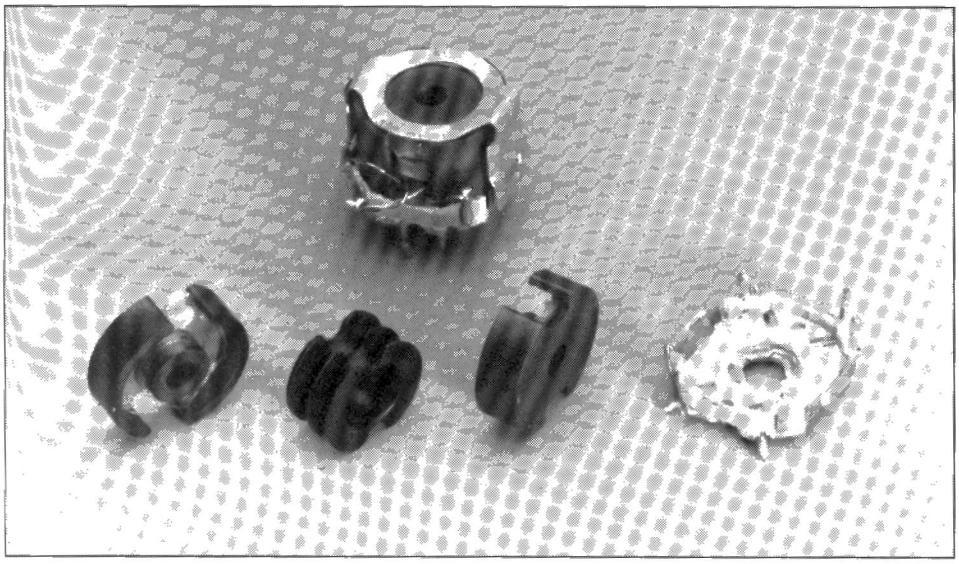


Fig. 11. Illustrating the construction of the special transformer.

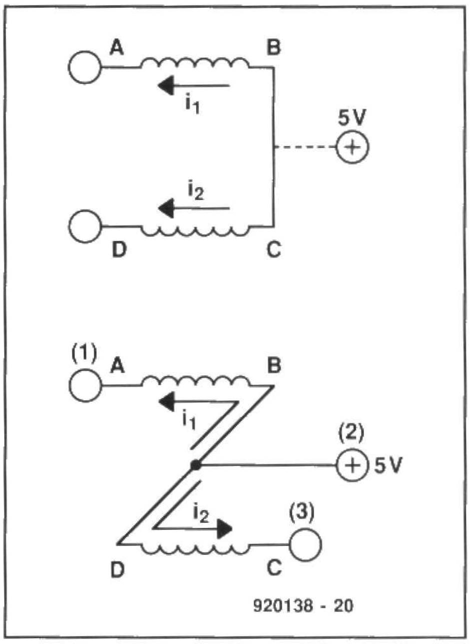


Fig. 12. Transformer windings and connections. Note how the two inductors are interconnected.

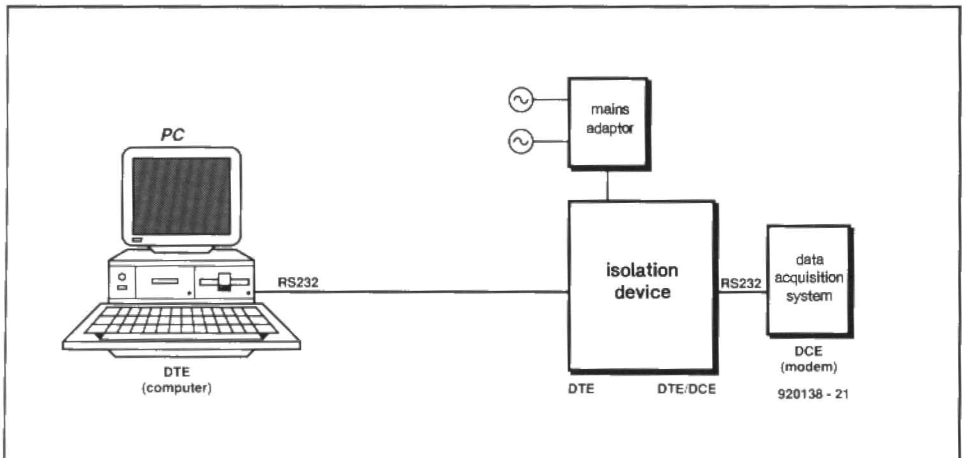


Fig. 13. Schematic representation of an automated test system. The RS232 isolator is inserted between the PC and the data acquisition device. For the best possible common-mode suppression, the isolator has to be located as close as possible to the modem.

former, and solder the wire ends to the pins on the transformer plate. The yoke connection of the pot core is not used (cut off the pin, or bend it away) because connecting it to one of the two grounds (both is, of course, out of the question) results in an increase of the stray capacitance between the primary and the secondary sides of the transformer (the increase is from 2.5 pF to 5 pF). The isolation afforded by the transformer is sufficient to prevent arcing between the primary and the secondary at alternating voltages of 220 V and more. ■

WORLD BAND RADIO

Listening to the short wave bands is a hobby which has enthralled people since the very first radio transmissions were made. Whether eavesdropping on radio amateurs, listening to broadcast stations, or monitoring aircraft and shipping, there is a unique fascination about listening to stations from all over the world.

By Ian Poole, G3YWX

RECENT advances in technology have opened up the short waves to many more people. Previously, portable radios with short wave bands were not easy to use except for casual listening. Now a new breed of radios has come onto the market. World Band Radios as they are often termed have overcome the shortcomings of the earlier sets, and make it possible to have a set with good performance for a reasonable price. As a result many more people can listen to all manner of transmissions from all the corners of the earth.

Facilities

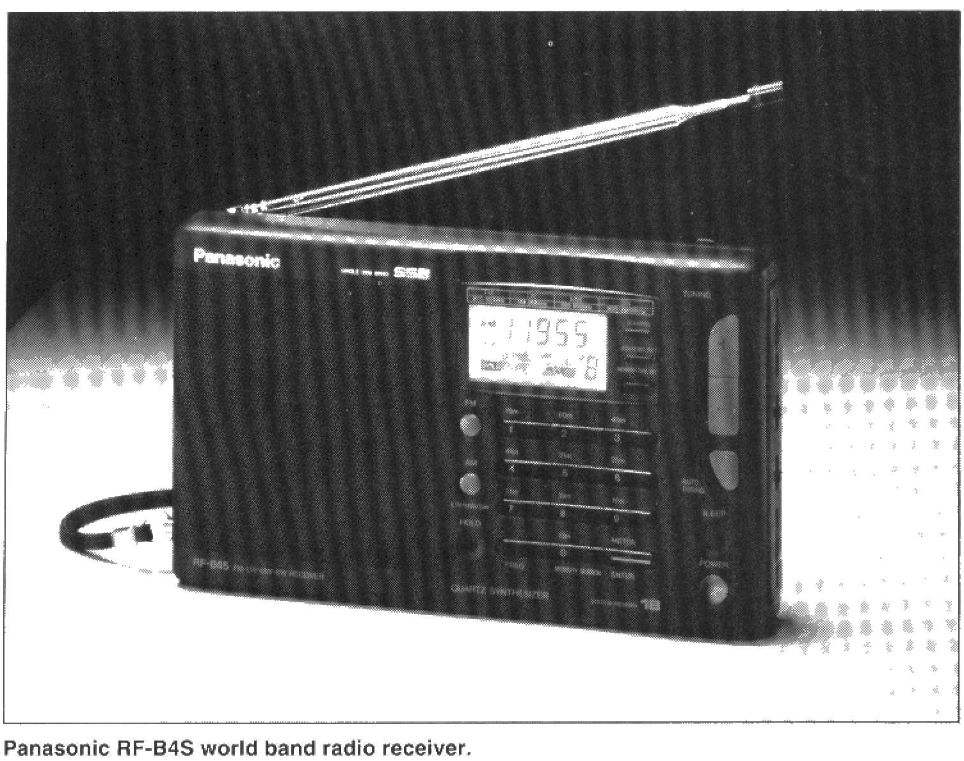
World Band Radios use some of the latest microprocessor and IC technology to give them a host of useful facilities. For example, tuning can be accomplished either by using a standard tuning knob, or by entering the frequency directly on to the keypad on the set. Furthermore, there is a digital readout of the exact frequency. This means that if the frequency of the station is known then the set can be tuned to it quickly and accurately.

Usually, these sets cover a wide band of frequencies. Typically, one might cover frequencies from as low as 150 kHz right up to 30 MHz. In addition to this, they usually cover the VHF FM band between 88 and 108 MHz. With this sort of coverage it is possible to listen to all the regular broadcast stations on the long and medium waves, as well as the transmissions on VHF FM. Whenever a change is required, frequencies in the short waves can be selected to hear different types of station.

A number of other facilities are included as well. There is a beat frequency oscillator (BFO) for the reception of morse and single sideband (SSB). Morse is still widely used and can be heard on many parts of the spectrum. Single sideband is a form of voice transmission which is widely used for long distance communications.

To make the fullest use of the microprocessor control, other features are included. For example, a clock is standard on most radios and this can even be used to turn the radio on or off, making it function as a clock radio. Memory channels are incorporated, making it very easy to select a favourite station. If this was not enough, many radios have a sweep facility included where the receiver will tune itself up or down in frequency, stopping when it detects a signal.

As these radios are designed specifically to perform over a wide range of frequencies and not just on the local broadcast bands, their RF performance is much better than the average transistor portable. They have better filters to ensure off-channel signals are suitably rejected. In addition to this, their general susceptibility to unwanted signals is much better. These factors make these radios much better for monitoring the short waves.



Panasonic RF-B4S world band radio receiver.

Using the radio

In view of the number of facilities which world band radios possess, they have many more controls than a normal domestic receiver. Fortunately, they are usually not difficult to use, although they may require a little familiarization.

Not all the controls will be new. There will be the normal volume and tuning knobs, and there will often be some tone controls. Some radios even have a balance control which comes into use when listening on headphones.

One of the most prominent sets of controls which marks out this type of set from more ordinary ones is the keypad. This is quite easy to use. Normally, the new frequency just has to be entered and then an enter key pressed. Other functions controlled by keypad including sweeping, memories and clock setting are also similarly easy to use.

The beat frequency oscillator (BFO) needs a little more explanation. It is used in the reception of morse and single sideband (SSB) signals. In most cases, a pitch control will be provided and it will have to be set slightly to one side of centre. The actual side will depend upon whether upper or lower sideband is being received. Normally, lower sideband is used below 10 MHz, and upper sideband

is used above this frequency. The settings for these should be marked on the set. For morse transmissions it normally makes little difference which side the control is set.

On some sets there may be a filter switch. This should be set to wide for most broadcast stations, but reduced to narrow when there is interference, or when SSB is being received.

An RF gain control may be included. This adjusts the gain of the early RF stages and will prevent the set overloading when strong signals are being received. It can usually be left at maximum when using the set's own aerials, but if an external aerial is used it may need to be reduced from its maximum setting.

Better reception

World band radios will be able to give quite good reception on their own. For short wave transmissions, the telescopic aerial should be extended as far as possible. Ideally, this should be kept away from any metalwork or wiring for the best signal strengths.

In addition to this there are many sources of interference within the average home these days. Electrical motors are a traditional source of noise. Fridge and freezer motors generally just give a large click or plop as they are turned on or off, but electric drill motors are notoriously noisy, generating high levels of broadband interference.

Televisions and computers are another source. The line scanning circuitry produces very high level signals which are radiated together with their harmonics and can be detected up to 10 MHz and higher. The computers themselves can also generate a lot of interference.

In order to improve reception, the radio should be placed as far away from any of these sources of interference as possible. This may not always be easy, and it is generally a compromise between convenience and optimum reception.

To make the best of the receiver, and obtain the best reception, an external aerial can be used. Most sets have a special socket at the back for this purpose, and they will also have a switch for selecting the internal or external aerial. This can be easily forgotten and must be in the correct position for the aerial which is in use.

For most applications a random length of wire as high and as long as possible is ideal. This type of aerial is commonly called a long-wire, although it is more correctly called an end-fed wire. If at all possible, it should be outside the house, as shown in Fig. 1, where it will give much better reception of the wanted signals and lower levels of interference from sources like televisions, computers and the like.

An aerial like that shown in Fig. 1 is likely to pick up signals from most directions particulary if the vertical section forms a large part of it. Whilst this is quite satisfactory in many instances, it can often be advantageous to use a directional aerial. By using one of these aerials it is usually possible to reduce the levels of interference. This is because the aerial can be rotated to give the strongest signal from the wanted station. As other interfering signals are likely to be coming from different direction they will be reduced in strength. Examples of home-made directional aerials are described in Refs. 1, 2 and 3.

Table 1. Designations of the radio spectrum

Frequency (MHz)	Designation
0.003 – 0.03	Very Low Frequency (VLF)
0.03 – 0.3	Low Frequency (LF)
0.3 – 3.0	Medium Frequency (MF)
3.0 – 30	High Frequency (HF)
30 – 300	Very High Frequency (VHF)
300 – 3000	Ultra High Frequency (UHF)
3000 – 30000	Super High Frequency (SHF)
30000 – 300000	Extra High Frequency (EHF)

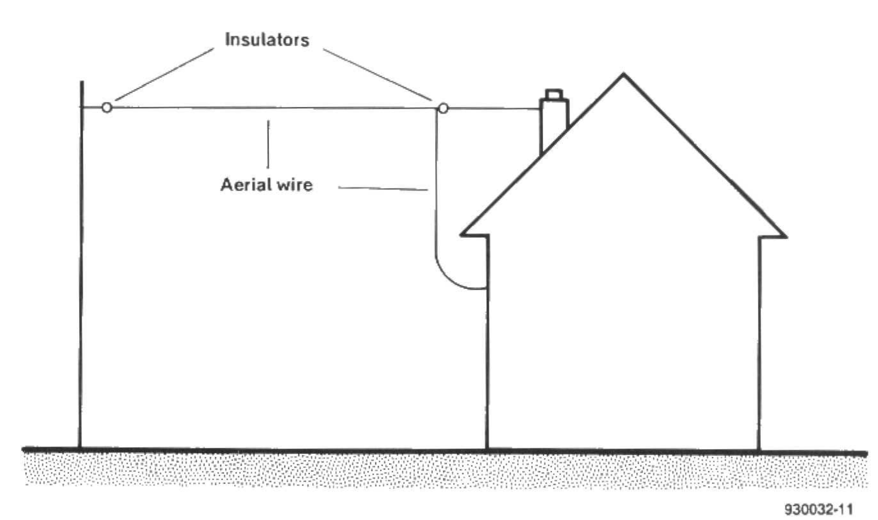


Fig. 1. A typical end-fed wire aerial construction.

Signal Propagation

Signals can travel for many thousands of miles on the short wave bands. In fact, it is not unusual to hear signals from the

other side of the globe at some times of day.

Like light, radio waves do not normally curve along the surface of the earth. To be able to reach the colossal distances around the earth's surface, the radio waves are reflected by an area of the atmosphere called the ionosphere. This stretches between about 100 and 400 km above the earth's surface.

In this area — and just below it — there are a number of different ionized layers as shown in Fig. 2. The lowest of these layers is the D layer, and it is only present during the day. It tends to absorb low-frequency signals. This is why signals do not travel very far during the day on the Medium Wave band, but at night they can be heard at much greater distances.

As the frequency rises, signals start to pass through the D layer and reach the E layer. Here the signals are not absorbed as much, but instead they are bent around so that they are reflected back to earth as shown in Fig. 3.

As the signal frequency increases still further, the bending or refraction is reduced. Eventually, a point comes when the signals pass through this layer and reach the next one called the F layer. This is the highest and also the most changeable of the ionized layers. During the day it splits into two layers, called the F1 and F2 layers, but at night they combine.

Again, signals reaching these layers will be bent back to earth. And again it is found that this refraction becomes less as the frequency increases and the signals eventually pass through, and travel on into outer space.

In general, the frequencies which are affected by the ionosphere are those below about 30 MHz. However, this is only a rough guide because the continually changing state of the various layers means that the different frequencies are affected. At certain times, it is possible for frequencies as high as 60 MHz or more to be affected. Other times, signals as low as 20 or 25 MHz pass through into space.

These reflections enable signals on the short waves to travel over vast distances. However, it does not fully explain how signals manage to travel from one side of the earth to the other. The reason lies in the fact that signals, once they have been reflected down by the ionosphere, may be reflected by the earth itself. In this way, radio signals can undergo several reflections and thereby reach any point around the world.

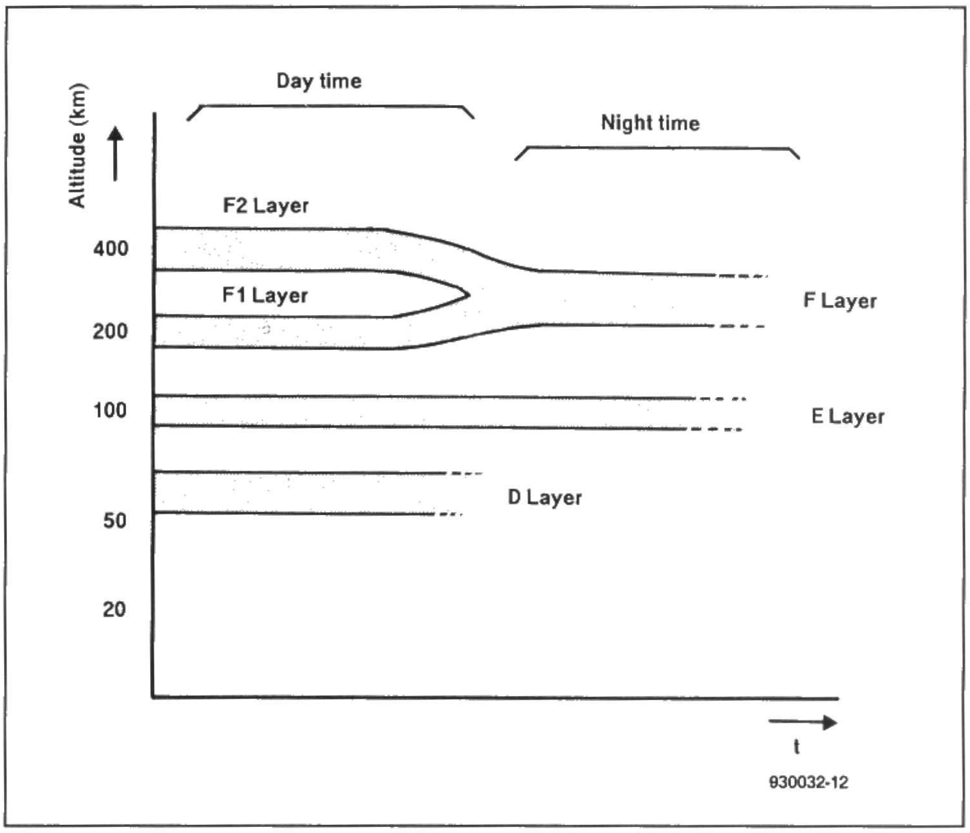


Fig. 2. The ionized layers around the earth.

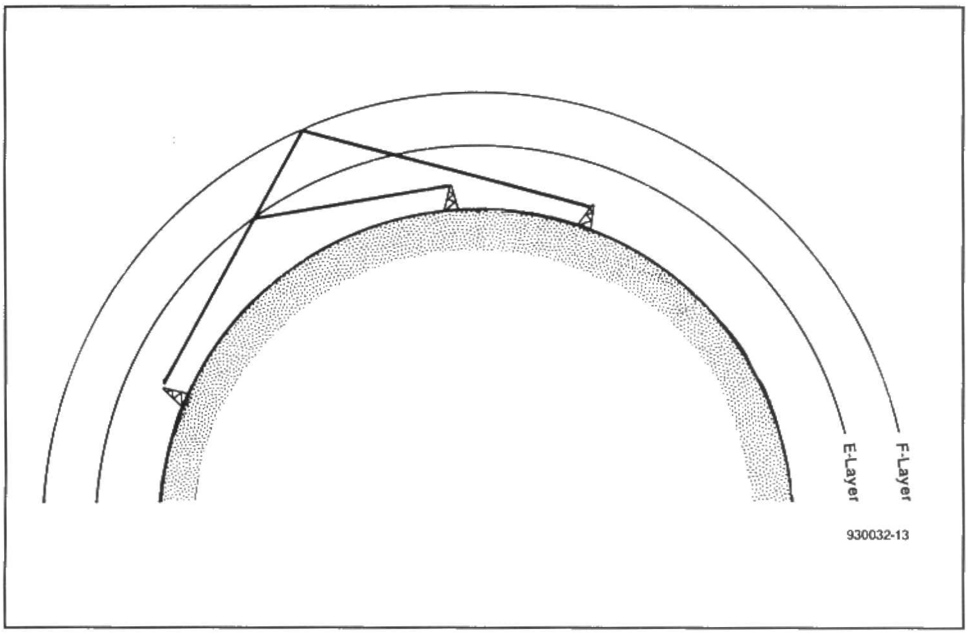


Fig. 3. Radio signals reflected by the ionosphere.

What can be heard

There is an enormous number of different organizations and interests which use the short wave bands. One of the most obvious is international broadcasting. Radio amateurs also make good use of these frequencies as well. In addition, there are a whole host of other users.

Ships rely heavily on radio for their communications. They use some of the lower frequencies just above the medium wave band for comparatively short range communications of up to a few hundred miles. In fact fishing boats use these frequencies a lot. Higher frequencies are used for much greater distances. Cargo ships many thousands of miles from their home ports use these frequencies to keep in contact with their companies and obtain instructions or give reports. However, with many of the current developments in technology, satellites operating well above the range of world band radios are being used increasingly.

Another important use for the short waves is for aircraft to maintain contact with the ground on long transatlantic trips. Generally, frequencies in the VHF portion of the spectrum are used when the aircraft is up to 50 miles or possibly more away from the air traffic control station. For communications of a few hundred miles or more, frequencies in the HF portion of the spectrum need to be used. Whilst satellites are also being used increasingly for this purpose, the

HF bands still carry most of this type of communications.

Apart from these users, there is a whole host of other organizations who use these frequencies. For instance, government departments including embassies. One only has to look at the top of any embassy to see the large array of aerials used to communicate with their home countries. In addition to this, the military make very good use of radio, as do a number of private concerns as well.

Bands

To accommodate all of these services in an efficient way, the radio spectrum is split up, and portions allocated to different users. In this way, the best use is made of the available spectrum, and the minimum amount of interference is caused. From Table 2 it can be seen that a number of bands are allocated to the broadcasting services, and amateurs have their own allocations as in Table 3. In this way, amateurs who are only allowed to use comparatively low powers can still operate without undue interference from broadcasting stations who use many kilowatts of power. Similarly, other users are given their portions of the radio frequency spectrum.

To co-ordinate the way in which the different frequency bands are allocated, international agreement is needed. An organization called the International Telecommunications Union (ITU) performs this role. Whilst it has an ongoing function, every few years large conferences called World Administrative Radio Conferences (WARC) are held. Here, the allocations for the different types of service are updated and agreed. The last one was held in early 1992 at Torremolinos in Spain. A number of changes were made to allocations — most were in the microwave region of the spectrum, although a number will affect the short wave bands.

Short wave broadcasting

Listening to short wave broadcast stations can be a fascinating hobby in itself. At one time or another it is possible to hear stations from all over the world, and with the number of stations filling the bands, there is a very wide selection to choose from.

Stations broadcast a wide selection of programmes. From news and politics to music and programmes about their countries, almost every taste is accommodated, and there is more than enough to prevent one from ever becoming bored.

Listening to news and political programmes can be particularly interesting. As stations are generally run by their governments, they often act as an official mouthpiece for their country. This means that it is not unusual to hear views or aspects to international news which are totally different to those heard on our domestic broadcasts.

Often, short wave broadcasts are used for propaganda purposes. Many Eastern European countries used to operate (and some still do) some very large and powerful transmitters. Radio Moscow and Radio Tirana (Albania) are probably two of the most famous. They could nearly always be heard on a host of different frequencies on several of the broadcast bands. More recently, the importance of short wave broadcasting was demonstrated during the Gulf war. Iraq used to have some powerful transmitters which it used for this purpose to great effect. Not surprisingly, the coalition forces retaliated by also setting up a number of transmitters.

There are a total of 13 different short wave broadcast bands, as shown in Table 2. They are situated throughout the short wave spectrum, so that the broadcasters can make the best use of the propagation characteristics as they change with frequency. This enables transmissions to be targeted to the correct part of the world.

Of the 13 bands, the 120, 90, and 60-metre bands are known as tropical bands. This is because they are only used in tropical areas of the world where noise levels caused by electrical storms make the conventional long and medium wavebands unusable.

Of the other bands, the 49 and 41-metre bands are probably the most popular amongst broadcasters. They are ideal because they give reliable medium-distance reception. This means that they are very crowded, but it is often possible to hear stations up to about 3000 km away. More distant weaker stations are often masked by the stronger more local ones. However, it is often possible to receive some long-distance stations at night with some careful listening.

For those wanting to hear long-distance stations bands, the 16 and 13-metre bands are good hunting grounds. They are best at dawn and dusk, although they give a good selection of stations during the day, but at night conditions are usually worse. The highest frequency band, 11-metres, is little used. The reason for this is that propagation at these frequencies is less reliable than on the lower ones.

Currently, broadcast stations transmit AM (amplitude modulation). Whilst this is not the most efficient mode to use, many short wave receivers cannot resolve some of the more efficient modes like SSB. However, to enable sufficient stations to transmit on these bands, stations are spaced 5 kHz apart unlike on the medium and long wave bands where they are 9 kHz apart. The 9-kHz spacing enables higher quality transmissions to be made and with lower amounts of interference.

Table 2. Broadcast allocations

Long, Medium and Short Wave Broadcast Bands

Band	From	To
Long Wave	0.150	0.285
Medium Wave	0.5265	1.6065
120 Metres	2.300	2.498
90 Metres	3.200	3.400
75 Metres	3.950	4.000
60 Metres	4.750	5.060
49 Metres	5.950	6.200
41 Metres	7.100	7.300
31 Metres	9.500	9.900
25 Metres	11.650	12.050
22 Metres	13.600	13.800
19 Metres	15.100	15.600
16 Metres	17.550	17.900
13 Metres	21.450	21.850
11 Metres	25.670	26.100

(all frequencies in MHz)

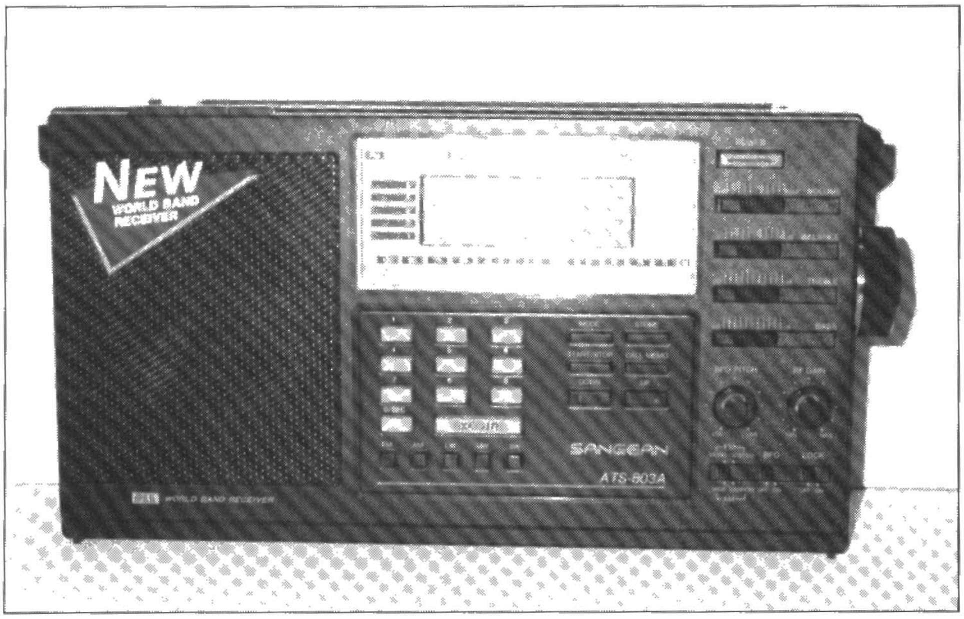
Table 3. UK SW amateur bands

Frequency limits (MHz)		Approximate wavelength
1.81	2.0	160 Metres (Top Band)
3.50	3.80	80 Metres
7.00	7.10	40 Metres
10.10	10.15	30 Metres
14.00	14.35	20 Metres
18.068	18.168	17 Metres
21.00	21.45	15 Metres
24.89	24.99	12 Metres
28.00	29.70	10 Metres

Radio amateurs

Radio amateurs, or 'hams', also make good use of the short wave bands. They have obtained licences which allow them to transmit to other radio amateurs. Often, they make contacts all over the world, many of them in remote and unusual places. This can make listening to them quite fascinating. In fact, part of the excitement of monitoring the amateur bands is that one never knows where the next station might be located. It could be in Europe, Russia, The States, or somewhere in the wilds of Africa, or a small island in the middle of the Pacific Ocean.

There are a number of amateur bands within the short wave part of the radio spectrum as shown in Table 3. Those lower in frequency (160 and 80 metres) are generally used for local contacts within the same country. However, at night time more distant contacts are possible. The 40 and 30-metre bands will usually give contacts around Europe by



Sangean ATS803A world band radio receiver.

day, and then at night distances increase. The higher frequency bands support intercontinental contacts most days. However, the two bands which are highest in frequency, ten and twelve metres, are the least reliable owing to the changes in the ionosphere.

A variety of different modes are used by amateurs. AM is seldom heard nowadays. Instead, for speech transmissions, single sideband is used. Morse is also widely used. It offers the advantage that it can be copied when signal strengths are low. Also, equipment for transmitting morse can be made more easily, a distinct advantage for those who construct their own equipment. Apart from these modes, amateurs are becoming increasingly interested in various types of data transmission. The original radio teleprinter transmissions of the 1950s and 60s have declined in popularity, giving way to new modes like Packet and Amtor which use computer technology for error checking and additional facilities.

For identification, each amateur station is allocated its own individual callsign. These callsigns consist of two parts. The first is the prefix, and from this it is possible to determine where the station is located by comparing the prefix with a list. The remainder of the callsign consists of up to three letters which act as serial figures. One example of an amateur callsign is G3YWX. G3 is the prefix, and indicates that the station is located in England. Some examples of amateur prefixes are shown in Table 4.

Whilst there are many other different groups of people and organizations which use the short wave bands, listening to broadcast stations and radio amateurs are probably the two main areas of interest. In fact, in the UK it is not legal to listen to most of the other transmissions which can be picked up.

Table 4.
Examples of amateur callsign prefixes

C3	Andorra
CE	Chile
DA-DL	Germany
EA	Spain
ES	Estonia
F	France
G	England
GI	Northern Ireland
GM	Scotland
GW	Wales
HA	Hungary
I	Italy
JA	Japan
K, KA-KZ	USA
LA	Norway
N, NA-NZ	USA
PA-PD	Netherlands
PY	Brazil
R, RA-RZ	C.I.S.
U, UA-UZ	C.I.S.
W, WA-WZ	USA
VK	Australia
ZL	New Zealand
2A-2Z	UK Novice licensees
4X, 4Z	Israel
5H	Kenya

Summary

For anyone interested in listening to the transmissions on the short waves, a large variety of radios is available. Manufacturers like Sony, Grundig and Panasonic produce a variety, ranging in price from a couple of hundred pounds upwards. For those interested in a less costly radio, the Sangean ATS 803A is available at just over £100 and gives excellent value for money. ■

References:
1. 'A compact spiral T/R HF antenna'. *Elektor Electronics* November 1992.
2. 'The flat-top 80 antenna'. *Elektor Electronics* March 1992.
3. 'Mark-2 80/40 QTC loop antenna'. *Elektor Electronics* July 1992.

Further reading:
1. World Radio and TV Handbook
2. An Introduction to Amateur Radio, by Ian Poole G3YWX, ISBN 0-85934-202-6. Publisher: Bernard Babani (publishing) Ltd. Price: £3.50.
3. An Introduction to Scanners and Scanning, by Ian Poole G3YWX, ISBN 0-85934-256-5. Publisher: Bernard Babani (publishing) Ltd. Price: £4.95.
4. Radio amateur's guide: Radio wave propagation (HF bands), by F.C. Judd G2BCX. ISBN 0-434-90926-2. Publisher: Heinemann Newnes.
5. 'Improving portable radio performance'. *Elektor Electronics* February 1992.

Additional information:
The Sangean ATS 803A is available from S.R.P.Trading, Unit 20, Nash Works, Forge Lane, Belbroughton, Nr Stourbridge, Worcestershire, England. Telephone (0562) 730672.

SIMPLE, LOW-COST ANTENNA TEST INSTRUMENTS - 1

This article looks at instruments and methods that can be used by both ham operators and shortwave listeners to get the most out of an antenna.

By Joseph J. Carr

MANY of the most popular, and most effective, radio antennas are resonant designs that must be properly adjusted for best operation. If these antennas are operated off resonance, their effectiveness is reduced considerably. Unfortunately, many of the antenna instruments that are easily available in hobbyist channels require a radio transmitter to energize them. For example, most standing wave ratio (VSWR) meters and nearly all RF power meters require transmitter power levels to operate; both can be used for making antenna adjustments. Amateur and professional radio transmitter operators use these instruments routinely. Unfortunately, receiver owners are not licensed to connect a transmitter to the antenna, so are banned from using the most common instruments. There are, however, several different instruments — that are easy to either build or buy — that can be used by shortwave and VHF/UHF scanner listeners, as well as ham operators, for adjusting or testing resonant antennas.

Before looking at specific instruments, let us first define the problem faced when either erecting an antenna for the first time, or testing an older antenna.

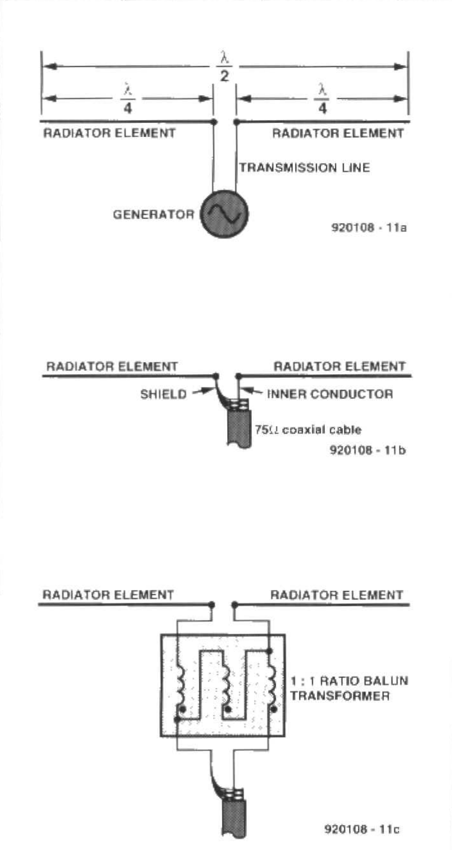


Fig. 1. Typical resonant antenna. This antenna is half wavelength dipole; b) coaxial cable fed dipole antenna; c) use of a 1:1 balun transformer at the feedpoint improves the dipole radiation and reception pattern.

The problem

The problem with resonant antennas can be seen by considering an example. Figure 1a shows the half wavelength dipole antenna. This type of antenna consists of two quarter wavelength legs fed at the centre by a transmission line. In the example of Fig. 1a, the dipole is fed with a generator, or transmitter, but because there is a law of reciprocal action in effect for antennas, a receiver could also be located at the feedpoint. The antenna performance is the same in either case. The direction of maximum radiation (or reception) in the horizontal plane (i.e., as seen from above) is perpendicular to the radiator elements, forming a 'figure-8' pattern. There are sharp nulls 'off the ends' in-line with the radiator elements.

Although shown here as a parallel line, it is more common to see coaxial cable transmission line (Fig. 1b). In many cases, the dipole is fed using coaxial cable and a 1:1 impedance ratio balun (BALanced UNbalanced) transformer (Fig. 1c). The use of a balun transformer makes it more likely that the radiation pattern will be somewhat like the idealized 'figure-8' pattern.

Figure 2 shows the electrical situation along the length of the antenna. The centre of a half wavelength radiator element is a current node and a voltage antimode, so the current (I) is maximum, and voltage (U) is minimum. The feedpoint impedance is minimum — ideally 73 Ω — and rises to about 2,500 Ω at the ends. This situation is found only at the frequency at which the antenna is resonant, so the antenna must be tuned for optimum performance. The way an antenna is tuned is to alter its length, i.e., trim or add equal amounts at both ends until the resonant point is found. It is customary to make the antenna longer than the design value, and then trim it for resonance; trimming is a bit easier than adding length.

Another problem is that the minimum impedance, which is 73 Ω for an ideal dipole in free space, varies from a very low value when the antenna is mounted close to the ground, to as much as about 120 Ω at critical heights.

Transmission lines

The transmission line is not merely a wire that carries RF power to the antenna. It is actually a complex circuit that simulates an infinite L-C network. There is a characteristic impedance (Z_0), also called surge impedance, that describes each transmission line. This impedance is the square root of the ratio of the capacitance and inductance per unit of length. When a load having a resistive impedance equal to the surge impedance of the transmission line is connected to the antenna, then the maximum transfer of power between the line and the antenna takes place.

We cannot deal extensively with transmission line theory here. However, it is prudent to have at least have some idea of what the circuit looks like. Figure 3 shows a model of a transmission line in which Z_0 is the characteristic impedance of the line, $R2$ is the load impedance of the antenna, and $R1$ is the output impedance of the transmitter. In a properly de-

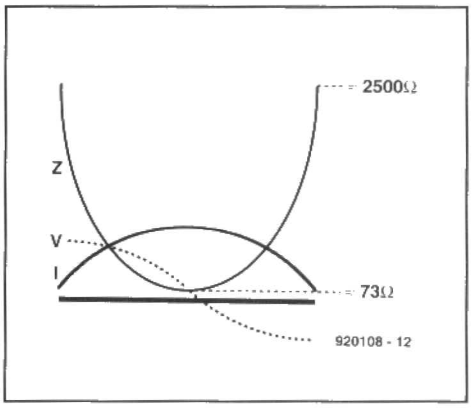


Fig. 2. Voltage (U), current (I) and impedance (Z) relationships along the length of a dipole antenna.

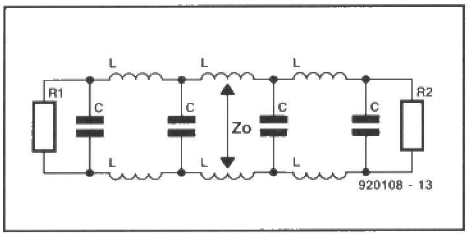


Fig. 3. Equivalent *L-C* circuit of a transmission line.

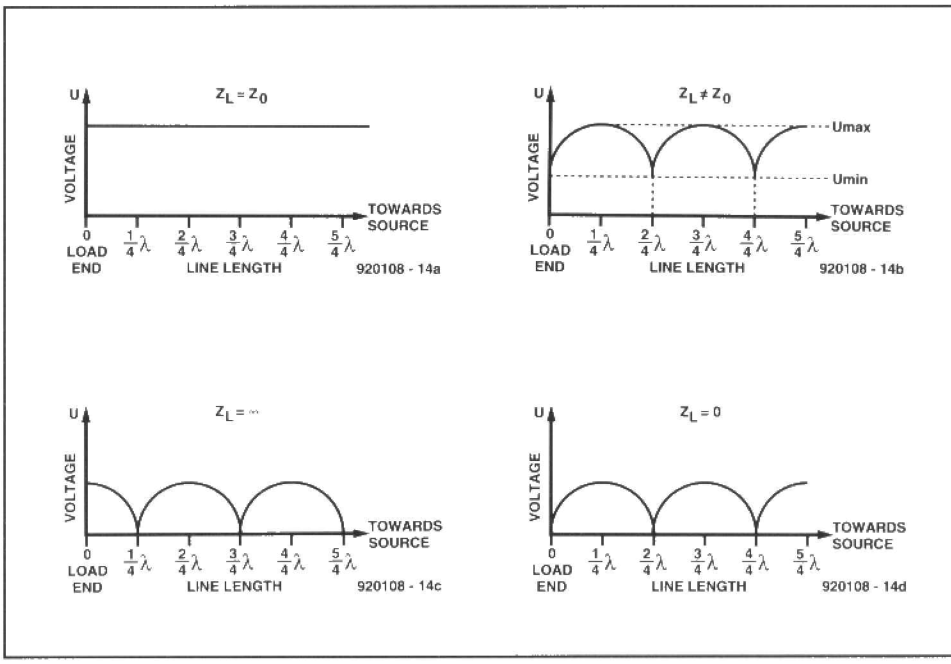


Fig. 4. Voltage along a transmission line under various circumstances: a) line flat ($Z_o = Z_L$); b) line and load impedances different ($Z_o \neq Z_L$); c) open line ($Z_L = \infty$); d) line shorted ($Z_L = 0$).

signed antenna system, all three impedances will be either equal ($Z_o = R1 = R2$), or a matching network will make them appear equal.

We must consider the electrical situation along the transmission line in order to understand the readings that we see on our instruments. Figure 4 shows several possible situations. These graphs are of the RF voltage along the line, with voltage on the vertical axis and transmission line length (expressed in wavelengths of the RF signal) along the horizontal axis. When the system is matched ($Z_o = R2$), the voltage is the same everywhere along the line (Fig. 4a). This line is said to be 'flat'. But when Z_o and $R2$ are not equal, then the voltage varies along the line with as a function of electrical length. In mismatched antenna systems not all of the power is radiated, but rather some of it is reflected back towards the transmitter. The forward and reflected waves combine algebraically at each point along the line to form standing waves (Fig. 4b). We can plot the voltage maxima (U_{max}) and minima (U_{min}). Keep this graph in mind because we will refer again to it when we deal with VSWR.

Two special situations can occur in transmission line and antenna systems that yield similar results. The entire forward power is reflected back to the transmitter (none radiated) if the load (i.e., antenna) end of the transmission line is either open or shorted. The voltage plot for an open transmission line ($R2$ is infinite) is shown in Fig. 4c, while that for the shorted line is shown in Fig. 4d. Note that they are very similar to each other except for where the minima ($U_{min} = 0$) occur. The minima are offset from each other by 90 degrees (i.e., quarter wavelength).

Calculating standing wave ratio

When a radio signal passes down the transmission line to the antenna, some of it is radiated, and some of it is reflected back towards the source. When the forward wave and the reflected wave interfere with each other, the standing waves are formed on the line. The standing wave ratio, usually referred to as the voltage standing wave ratio or VSWR, is a measure of how well the antenna is impedance matched to the transmission line. It can be used as an indicator of resonance because the VSWR is minimum, i.e., 1:1, when the antenna is both resonant and matched to the transmission line.

The VSWR can be calculated from any of several bits of knowledge. Even if you do not have a VSWR meter, therefore, it is possible to determine VSWR. If the antenna load impedance ($R2$) is not equal to Z_0, then we can calculate VSWR from one of the following:

If Z_0 is greater than $R2$:

$$\mathrm{VSWR} = Z_0 / R2 \qquad (1)$$

If Z_0 is smaller than $R2$:

$$\mathrm{VSWR} = R_2 / Z_0 \qquad (2)$$

We can also measure the forward and reflected power, and calculate the VSWR from those readings:

$$\mathrm{VSWR} = \frac{1+\sqrt{P_r / P_f}}{1-\sqrt{P_r / P_f}} \qquad (3)$$

Where:
VSWR is the voltage standing wave ratio;
P_r is the reflected power;
P_f is the forward power.

If we can measure either the voltage maxima and minima, or the current maxima and minima, then we can calculate SWR:

$$\mathrm{VSWR} = \frac{U_{max}}{U_{min}} = \frac{I_{max}}{I_{min}} \qquad (4)$$

Finally, if the forward and reflected voltage components at any given point on the transmission line can be measured, then we can calculate the VSWR from:

$$\mathrm{VSWR} = \frac{U_f + U_r}{U_f - U_r} \qquad (5)$$

Where:
U_f is the forward voltage component;
U_r is the reflected voltage component.

The latter equation, based on the forward and reflected voltages, is the basis for many modern VSWR and RF power meters.

The goals of antenna instrumentation

There are three different goals that lead to the same place when using antenna instruments to make adjustments. One goal is to find the resonant frequency of the antenna, regardless of the feedpoint impedance. Second, we can measure the feedpoint impedance of the antenna at a frequency in the centre of the band of interest, and then select a transmission line with a characteristic impedance that matches it. For example, when adjusting a dipole, we can measure the feedpoint impedance at the design frequency, and then adjust the antenna dimensions to provide the closest match to 73 Ω. Third, one can implicitly measure the feedpoint impedance by measuring the voltage standing wave ratio (VSWR). The VSWR is an indication of the ratio of the transmission line characteristic impedance (Z_o) and the feedpoint impedance of the antenna (Z_L).

The goal when adjusting a resonant antenna is two-fold. First, the length must be adjusted to resonance in the middle of the band of interest. For example, if you are interested in the 25-metre broadcast band, you might want to tune the antenna length to 11,750 kHz. Resonance is found by finding the frequency at which minimum VSWR ap-

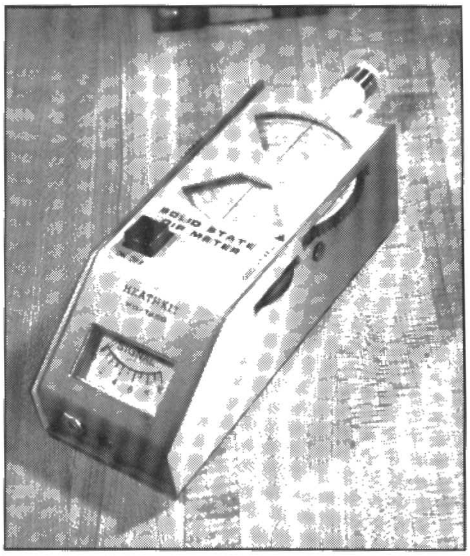


Fig. 5. Dip oscillator built from a kit.

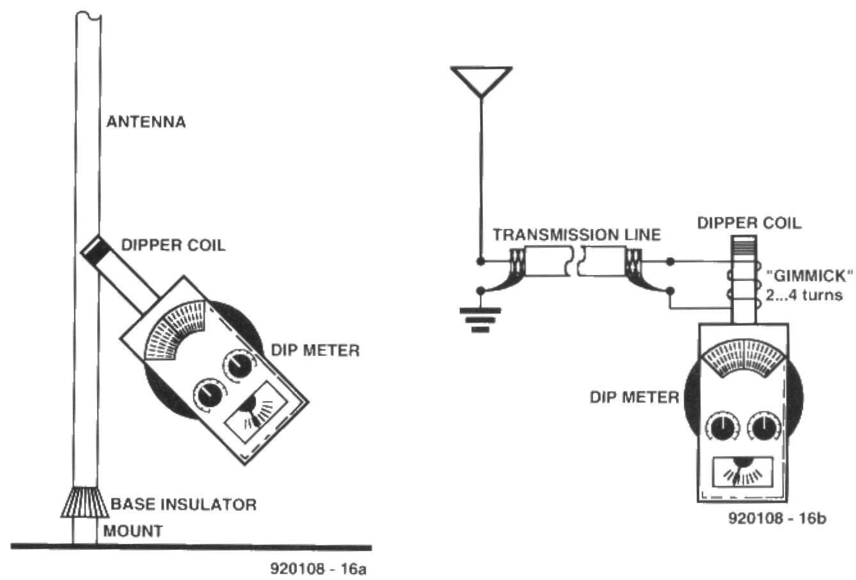


Fig. 7. a) Coupling a dip oscillator to a vertical antenna; b) coupling a dip oscillator to an antenna through a transmission line.

pears. When adjusting the antenna, the desired frequency is used, and the antenna length adjusted to minimize VSWR. The second goal is to measure the feedpoint impedance, or at least its resistive component, in order to provide impedance matching through an antenna tuner or other methods, if needed.

Several different instruments will achieve these goals; dip oscillators; impedance bridges; noise bridges; and self-contained VSWR analyzers. We will look at each category of instruments.

Dip oscillators

One of the most common instruments for determining the resonant frequency of an antenna is the so-called dip oscillator, or dip meter. Originally called the grid dip meter, the basis for this instrument is the fact that its output energy can be absorbed by a nearby resonant circuit (or antenna, which electrically is the same as a resonant *L-C* tank circuit). A typical dip oscillator consists of an *L-C* tuned variable frequency oscillator (VFO) constructed such that the inductor in the VFO tuned circuit is external to the cabinet, and a frequency calibrated dial is ganged to the variable capacitor. In most dip oscillators, there are a number of coils, one each for the different bands of operation, that can be interchanged.

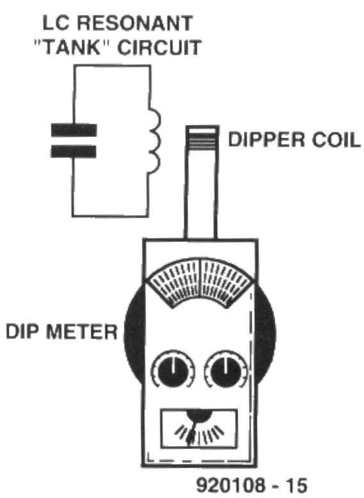


Fig. 6. Coupling a dip oscillator to an *L-C* tuned circuit.

Figure 5 shows a commercial dip meter built from a kit. When the inductor of the dip oscillator (see Fig. 6) is brought into close proximity to a resonant tank circuit, and the oscillator is operating on the resonant frequency, a small amount of energy is transferred. This energy loss shows up on the meter pointer as a slight 'dipping' action. The dip is extremely sharp, and is easily missed if the meter frequency dial is tuned too rapidly. Tune extremely slowly or you will probably miss the dip.

Antennas are resonant circuits, and can be treated in a manner similar to *L-C* tank circuits. Figure 7a shows one way to couple the dip oscillator to a vertical antenna radiator. The inductor of the dipper is brought into close proximity to the base of the radiator. In Fig. 7b we see the means for coupling dip oscillators to systems where the radiator is not easily accessed (as when the antenna is still erected). We connect a small two or three turn loop to the transmitter end of the transmission line, and then bring the inductor of the dipper close to it. A better way is to connect the loop directly to the antenna feedpoint.

There are two problems with dip meters that must be recognized in order to best use the instrument. First, the dip is very sharp. It is easy to tune past the dip and not even see it. To make matters worse, it is normal for the meter reading to drop off gradually from one end of the tuning range to the other; many beginners erroneously take this drop-off as the dip. But if you tune very slowly, you will notice a very sharp dip when the resonant point is reached.

The second problem is the dial calibration. The dial gradations of inexpensive dip meters are too close together, and are often erroneous. It is better to monitor the output of the dip oscillator on a receiver, and depend upon the calibration of the receiver for data. Keep in mind that the receiver must be tuned to the dip oscillator signal while the dipper is still engaged with the resonant tuned circuit or antenna. The simple oscillator circuits used in these instruments tend to 'pull' somewhat in operating frequency when coupled to an external circuit, so will be on a different frequency if checked when disengaged.

A note of caution is in order for those people who obtain surplus or used radio instruments. A number of vacuum tube based grid dip oscillators are available on the used and hamfest markets. These instruments are of very good design except for one problem: they connect to the AC power mains. I have seen instruments that are positively dangerous, partly from wear and partly from poor initial design. The problem is exacerbated by the fact that dippers are often used out of doors, where operator contact with the ground is highly probable. Replace the AC line cord with a modern line cord approved for outdoor use. In the USA, where I live, this means a three-wire cord in which the green or green/yellow wire is grounded to the cabinet and chassis of the instrument. As a prudent safety measure, never operate AC mains powered equipment out of doors without using an 1:1 isolation transformer approved for the purpose. ❑

Continued next month

PRODUCT OVERVIEW

MARCH 1993

ELEKTOR ELECTRONICS

A number of projects carried in *Elektor Electronics* are supported by ready-made printed-circuit boards (PCBs), self-adhesive front panel foils, ROMs, EPROMs, PALs, GALs, microcontrollers and diskettes, which may be ordered through our Readers Services using the order form printed every month opposite the Readers Services page.
The list printed here is complementary to the shorter one opposite the Readers Services page elsewhere in this issue. This two-page overview of all currently available products is regularly updated and will appear in the March, June, September and December issues of *Elektor Electronics*.
Items marked with a dot (•) following the product number are in limited supply only, and their availability can not be guaranteed by the time your order is received.
Items not listed here or on this month's Readers Services page are not available.
The artwork for making PCBs which are not available ready-made through the Readers Services may be found in the relevant article (from March 1990 onwards).
Prices and item descriptions subject to change. Prices can be confirmed on request at the time of ordering. Sterling (£) prices are inclusive of VAT at 17.5%.

PRINTED-CIRCUIT BOARDS

1986

Project	No.	Price (£)	(US$)
APRIL 1986			
Portable mixer	86012-1 •	6.25	12.50
MAY 1986			
Printer buffer	85114-1 •	13.80	27.60
	85114-2 •	5.95	11.90
JUNE 1986			
Rain gauge	86068 •	4.25	8.50
SEPTEMBER 1986			
RTTY interface	86019 •	8.95	17.90
Universal peripheral equipment	86090-1 •	9.35	18.70
OCTOBER 1986			
IDU for satellite TV reception	86082-1 •	14.80	29.60
Computerscope	86083 •	28.90	57.80
	9968-5	2.40	4.80
NOVEMBER 1986			
Top of the range preamplifier	86111-3A	8.10	16.20
	86111-1	12.20	24.40
DECEMBER 1986			
Temperature probe for DMM	86022 •	1.25	2.50

1987

Project	No.	Price (£)	(US$)
JANUARY 1987			
Top-of-the-range preamplifier	86111-2 •	26.45	52.90
FEBRUARY 1987			
Electron ROM card	86089 •	6.70	13.40
MARCH 1987			
MSX EPROMmer	87002 •	11.15	22.30
Valve preamplifier (1)	87006-1 •	10.00	20.00
	86111-3A •	8.10	16.20
APRIL 1987			
Valve preamplifier (2)	87006-2 •	14.70	29.40
Facsimile interface	87038 •	10.40	20.80
MAY 1987			
MIDI signal distribution	87012 •	8.70	17.40
JULY/AUGUST 1987			
Headphone amplifier	87512 •	10.60	21.20
OCTOBER 1987			
Low-noise microphone preamplifier	87058 •	4.05	8.10
NOVEMBER 1987			
SSB receiver for 80m and 20m	87051 •	17.35	34.70
BASIC computer	87192	23.80	47.60
Dimmer for inductive loads	87181 •	7.05	14.10

1988

Project	No.	Price (£)	(US$)
JANUARY 1988			
Stereo limiter	87168 •	8.50	17.00
Switch-mode PSU	880001 •	5.90	11.80
FEBRUARY 1988			
Infra-red headphones	87640 •	7.20	14.40
MARCH 1988			
Computer-controlled slide fader	87259 •	18.80	37.60
Low-noise preamplifier for FM receivers	880041 •	7.65	15.30
Signal divider for satellite TV receivers	880067 •	5.90	11.80
APRIL 1988			
Fuzz unit for guitars	87255 •	7.65	15.30
Active loudspeaker system	880030 •	8.80	17.60
MAY 1988			
Plotter	87167 •	13.50	27.00
Balanced line driver and receiver	87197 •	12.35	24.70
JUNE 1988			
Wideband active aerial for SW receivers	880043-1 •	7.05	14.10
	880043-2 •	5.60	11.20
HF operation of fluorescent tubes	880085 •	11.45	22.90
JULY/AUGUST 1988			
I/O extension card for IBM PCs	880038	33.60	67.20
Frequency read-out for SW receivers	880039 •	21.60	43.20
Simple 80m RTTY receiver	886034X •	9.60	19.20
OCTOBER 1988			
Centronics interface for slide fader	880111 •	9.10	18.20
Preamplifier for purists	880132-1 •	6.95	13.90
	880132-2 •	14.40	28.80
Peripheral modules for BASIC computer	880159 •	5.90	11.80
NOVEMBER 1988			
Bus interface for hi-res LCD screens	880074 •	19.70	39.40
LFA-150 — a fast power amplifier	880092-1 •	9.95	19.90
	880092-2 •	9.20	18.40
Harmonic enhancer	880167 •	7.40	14.80
Portable MIDI keyboard	880168 •	9.20	18.40
DECEMBER 1988			
LFA-150 — a fast power amplifier	880092-3 •	7.50	15.00
	880092-4 •	7.60	15.20
CVBS-to-TTL adaptor	880098 •	5.70	11.40
Colour test pattern generator	880130 •	15.65	31.30
Autonomous I/O controller	880163	5.70	11.40
	880184 •	18.00	36.00
Pitch control for CD players	880165 •	13.50	27.00

1989

Project	No.	Price (£)	(US$)
JANUARY 1989			
Fax interface for Atari ST and Archimedes	880109	8.65	17.30
MIDI control unit	880178-1	10.65	21.30
	880178-2	7.80	15.60
Low-budget capacitance meter	UPBS-1	2.30	4.60
FEBRUARY 1989			
Digital Model Train	87291-1	4.95	9.90
Touch key organ	886077	11.80	23.60
VHF receiver	886127	8.75	17.50
MARCH 1989			
Power line modem	880189	7.15	14.30
Centronics buffer	890007-1	23.05	46.10
	890007-2	2.55	5.10
	890007-3	9.80	19.60
APRIL 1989			
Digital Model Train	87291-2/3	5.05	10.10
Function generator	UPBS-1	2.30	4.60
Triplet	890013-1	7.80	15.60
	890013-2	8.00	16.00
Multi-point IR control	890019-1	4.05	8.10
	890019-2	4.75	9.50
Video recording amplifier	ELV project		
MAY 1989			
RDS decoder	880209 •	5.30	10.60
Digital Model Train (4)	87291-4	6.15	12.30
Analogue multimeter	890035	14.70	29.40
DTMF system decoder	890060	7.65	15.30
Sine-wave converter	UPBS-1	2.30	4.60
S-VHS-to-RGB converter	ELV project		
JUNE 1989			
8-digit frequency meter	880128	13.50	27.00
Echo unit	ELV project		
JULY/AUGUST 1989			
MIDI keyboard interface decoder board	890105-1	8.25	16.50
controller board			
Tracking tester	ELV project		
Floppy disk monitor	890078	5.00	10.00
Function generator	UPBS-1	2.30	4.60
SEPTEMBER 1989			
Digital Model Train	87291-6	7.85	15.70
Stereo viewer	890044	6.55	13.10
Centronics monitor	890123	6.45	12.90
Resonance meter	886071	4.60	9.20
OCTOBER 1989			
Logic analyser with Atari ST	890126	6.15	12.30
CD error detector	890131	7.05	14.10
RGB-to-CVBS converter	ELV project		
NOVEMBER 1989			
Digital Model Train (8)	87291-5	51.10	102.20
DECEMBER 1989			
Digital Model Train	87291-7	10.30	20.60
EPROM simulator	890166	11.75	23.50
Hard disk monitor	890186	12.95	25.90
IC tester	ELV project		
LF/HF signal tracer	890183	9.40	18.80
Solid-state preamp	890170-1*	13.80	27.60
	890170-3*	10.60	21.20
Transistor curve tracer	890177	6.75	13.50

1990

Project	No.	Price (£)	(US$)
JANUARY 1990			
Video mixer (1)	87304-1	32.00	64.00
Mini EPROM programmer	890164	8.25	16.50
All solid-state preamplifier	890170-2*	18.50	37.00
Simple AC millivoltmeter	900004 •	7.65	15.30
1 GHz frequency meter card for PCs	894110	25.55	51.10

*The four PCBs required for the preamplifier (2 × 890170-1; 1× 890170-2 and 1× 890170-3) are available as a package, ref. 890170-9, at a discounted price of £48.15 (US$96.30).

Project	No.	Price (£)	(US$)
FEBRUARY 1990			
Initialisation aid for printers	900007	16.75	33.50
Digital Model Train (11)	87291-8	5.30	10.60
Reflex MW AM receiver	UPBS-1	2.30	4.60
Video mixer (2)	87304-2	19.10	38.20
Capacitance meter	900012 •	8.50	17.00
MARCH 1990			
Audio/video modulator	ELV project		
Digital model train (12)	87291-9	4.10	8.20
IC monitor	896140	8.80	17.60
Power line monitor	900025 •	5.60	11.20
Replacement for TCA280A	894078 •	6.45	12.90
Video mixer (3)	87304-3	41.70	83.40
APRIL 1990			
BBD sound effects unit	900010 •	9.10	18.20
Digital model train (13)	87291-10	4.70	9.40
Q meter	900031 •	7.05	14.10
RS-232 splitter	900017-1	8.50	17.00
	900017-2	5.30	10.60
Video line selector	900032 •	7.65	15.30
Wiring allocation tester	ELV project		
MAY 1990			
Acoustic temperature monitor	UPBS-1	2.30	4.60
Budget sweep/function generator	900040 •	8.25	16.50
Centronics ADC/DAC	900037D	17.90	35.80
PC servicing card	ELV project		
Transistor characteristic plotting	900058	5.60	11.20
JUNE 1990			
Electronic load simulator	900042 •	14.10	28.20
MIDI master keyboard	Doepfer Elektronik		
Mini EPROM viewer	900030	21.15	42.30
Power zener diode	UPBS-1	2.30	4.60
Remotely controlled stroboscope	ELV project		
JULY/AUGUST 1990			
Battery tester	ELV project		
Compact 10A power supply	900045	13.50	27.00
Intermediate projects	UPBS-1	2.30	4.60
Mini FM transmitter*	896118	5.00	10.00
Sound demodulator for satellite-TV receivers	900057	4.40	8.80
Audio power indicator	904004 •	4.40	8.80
Four-monitor driver for PCs	904067 •	6.15	12.30

* can not be supplied to readers in the UK

Project	No.	Price (£)	(US$)
SEPTEMBER 1990			
High-current h_{FE} tester	900078 •	6.45	12.90
Infra-red remote control	904085/86	7.95	15.90
Sound generator	ELV project		
OCTOBER 1990			
µP-controlled telephone exchange	900081	21.15	42.30
S-VHS/CVBS-to-RGB convertor	900055 •	14.40	28.80
NOVEMBER 1990			
400-watt laboratory PSU	900082 •	12.95	25.90
Active mini subwoofer	900122-1 •	7.05	14.10
Dubbing mixer EV7000	ELV project		
Medium-power audio amplifier	900098	10.60	21.20
Programmer for the 8751	900100	8.25	16.50
PT100 thermometer	900106 •	5.90	11.80
DECEMBER 1990			
Active mini subwoofer	900122-2 •	6.15	12.30
Millionmmeter	910004 •	5.90	11.80
Phase check for audio systems	900114-1/2 •	9.40	18.80
PC-controlled Video-text decoder (1)	ELV project		
Signal suppressor for all-solid state preamp	904024 •	4.40	8.80

1991

Project	No.	Price (£)	(US$)
JANUARY 1991			
Logic analyser (1):			
- Busboard	900094-4 •	10.60	21.20
PC controlled Video-text decoder (2)	ELV project		
SWR meter	900013	3.55	7.10
FEBRUARY 1991			
Logic analyser (2):			
- RAM board	900094-2 •	18.50	37.00
- Probe board	900094-3 •	5.00	10.00
Multifunction measurement card for PCs	900124-1	28.20	56.40
MIDI-to-CV interface	Doepfer Elektronik		
RDS decoder:			
- demodulator board	880209 •	5.30	10.60

Project	No.	Price (£)	(US$)
- processor board	900060 •	7.65	15.30
MARCH 1991			
The complete preamplifier:			
- input board	890169-1	26.10	52.20
- main board	890169-2	39.35	78.70
Electronic exposure timer	900041 •	10.85	21.70
PC-controlled weather station (1)	900124-3	4.40	8.80
2-m band converter	900006-1	5.00	10.00
APRIL 1991			
Logic analyser (3):			
- control board	900094-5 •	18.50	37.00
MIDI programme changer	900138 •	6.75	13.50
8-bit I/O for Atari	910005	12.35	24.70
6-m band transverter	910010	11.45	22.90
Wattmeter:			
- meter board	910011-1	6.45	12.90
- display board	910011-2	4.10	8.20
Moving-coil (MC) preamplifier	910016 •	10.60	21.20
Dimmer for halogen lights:			
- transmitter	910032-1 •	4.10	8.20
- receiver	910032-2 •	4.40	8.80
PC-controlled semi-conductor tester	ELV project		
MAY 1991			
80C32/8052 Computer	910042	12.05	24.10
Battery tester	906056	4.10	8.20
Laser (1)	ELV project		
Moving-magnet (MM) preamplifier	900111 •	6.75	13.50
Universal I/O interface for IBM PCs	910046	10.85	21.70
JUNE 1991			
Universal battery charger	900134	9.40	18.80
Logic analyser - 4			
- power supply board	900094-7 •	8.80	17.60
- Atari interface	900094-6 •	12.65	25.30
- IBM interface	900094-1 •	14.40	28.80
Digital phase meter (set of 3 PCBs)	910045-1/2/3	26.15	52.30
Light transceiver	UPBS-1	2.30	4.60
Variable AC PSU	900104 •	6.15	12.30
Light switch w. TV IR r/c	910048	5.60	11.20
RTC for Atari ST	910006	6.15	12.30
Stepper motor board -1:			
- PC insertion card	910054-1	29.10	58.20
JULY/AUGUST 1991			
Multifunction I/O for PCs	910029	24.40	48.80
B/W video digitizer	910053	22.60	45.20
Stepper motor board - 2:			
- power driver board	910054-2	28.50	57.00
Laser - 3	ELV project		
LED voltmeter	914005 •	5.60	11.20
Wien bridge	914007 •	4.10	8.20
Angled bus extension card for PCs	914030 •	12.05	24.10
Sync separator	914077 •	4.40	8.80
SEPTEMBER 1991			
Peak indicator for loud-speakers	ELV project		
Timecode interface for slide control:			
- main board	910055	24.40	48.80
- display board	87291-9a	4.10	8.20
Asymm-symm converter	910072	5.60	11.20
OCTOBER 1991			
PC-controlled weather station (2)	900124-2	3.80	7.60
Digital function generator			
- main board	910077-1	21.75	43.50
- display board	910077-2	12.65	25.30
Audio spectrum shift encoder/decoder	910105	10.35	20.70
NOVEMBER 1991			
Relay card for universal I/O interface	910038	12.95	25.90
Dissipation limiter	910071	4.40	8.80
Digital function generator			
- sine converter	910077-3	15.00	30.00
- R/T converter	910077-4	12.35	24.70
Class-A power amplifier (1):			
	880092-1	9.95	19.90
	880092-2	9.05	18.10
Timer for CH systems	UPBS-2	3.80	7.60
DECEMBER 1991			
Class-A power amplifier (2):			
	880092-3	7.50	15.00
	880092-4	7.60	15.20
Economy power supply	910111 •	9.40	18.80
µP programmable filters	910125	6.75	13.50
Amiga mouse/joystick switch	914078	4.10	8.20
A musical Christmas present	910157	3.25	6.50
Safe solid-state relay	914008 •	3.80	7.60
Slave mains on/off control Mark-2	914072 •	6.45	12.90
Wideband antenna amplifier	ELV project		

1992

JANUARY 1992

Project	No.	Price (£)	(US$)
CD player	910146	8.25	16.50
Fast precise thermometer	910081	8.50	17.00
Low-frequency counter			
- input board	910149-1	5.00	10.00
- display board	910149-2	6.45	12.90
Mini Z80 system	910060	10.60	21.20
Prototyping board for IBM PCs	910049	21.15	42.30
Universal RC5 code infra-red receiver	910137	4.70	9.40
PC-controlled weather station (3)	900124-5	10.00	20.00
FEBRUARY 1992			
Audio/video switching unit	910130	11.75	23.50
I²C interface for PCs	910131-1	14.40	28.80
Measurement amplifier	910144	13.50	27.00
Mini square wave generator	910151	5.30	10.60
RAM extension for mini Z80 system	910073	2.35	4.70
Switch-mode power supply	920001	4.40	8.80
MARCH 1992			
8751 emulator	920019	12.05	24.10
A-D/D-A and I/O for I²C bus	910131-2	6.15	12.30
AF drive indicator	920016	5.60	11.20
Centronics line booster	910133	5.90	11.80
FM tuner (tuner board)	920005	21.15	42.30
LC meter	920012	8.80	17.60
MIDI optical link	920014	6.15	12.30
APRIL 1992			
80C32 SBC extension	910109	13.50	27.00
2-metre FM receiver	910134	10.30	20.60
Comb generator	920003	8.50	17.00
AD232 converter	920010	12.35	24.70
Automatic NiCd charger	UPBS-1	2.30	4.60
LCD for L-C meter	920018	4.70	9.40
Milli-ohm meter adaptor	920020	4.40	8.80
MAY 1992			
1.3-GHz prescaler	914059	5.00	10.00
Compact mains supply	920021	7.35	14.70
FM tuner - 3 (PSU)	920005-2	8.80	17.60
GAL programmer	920030	11.15	22.30
NICAM decoder	920035	15.00	30.00
JUNE 1992			
4-Megabyte printer buffer	910110	18.80	37.60
Audio-video processor - 2	ELV project		
I2C display	920004	4.70	9.40
FM tuner - 4:			
- mode control board	920005-3	5.60	11.20
- synthesizer board	920005-5	10.85	21.70
Guitar tuner	920033	10.00	20.00
Multi-purpose Z80 card	920002	20.25	40.50
JULY 1992			
12VDC to 240VAC inverter			
- main board	920039-1	11.15	22.30
- power board	920039-2	6.45	12.90
Audio DAC - 1	920063-1	8.50	17.00
Optocard for universal PC I/O bus	910040	12.95	25.90
Audio-video processor - 3	ELV project		
FM tuner - 5:			
- keyboard/display	920005-4	14.40	28.80
- S-meter	920005-6	3.80	7.60
RS232 quick tester	920037	5.00	10.00
Small projects:			
Water pump control for solar power system	924007	7.35	14.70
Simple power supply	924024	5.00	10.00
Wideband active telescopic antenna	924102	3.25	6.50
SEPTEMBER 1992			
EPROM emulator - II	910082	10.00	20.00
Audio/video processor - 4	ELV project		
Audio DAC - 2	920063-2	18.80	37.60
OCTOBER 1992			
Audio DAC - 3	920063-3	26.45	52.90
Mains sequencer	920013	17.35	34.70
Wideband active antenna	924101	3.25	6.50
RDS demodulator	880209 •	5.30	10.60
NOVEMBER 1992			
Printer sharing unit	920011	14.70	29.40
Sound sampler for Amiga	920074	6.75	13.50
Difference thermometer	920078	5.30	10.60
Low-power TTL-to-RS232 interface	920127	3.55	7.10

For December 1992 to March 1993 items see page 70 of this issue.

ROMS — EPROMS — PALS — GALS — MICROCONTROLLERS

For pre-1989 project EPROMs see the December 1992 Product Overview or contact our Dorchester office

Article/Project	Issue	Order code	Description	Price (£)	(US$)
Multifunction measurement card for PCs	2/91	561	1 × 16L8	10.30	20.60
MIDI control unit	1/89	570	1 × 27C64	11.75	23.50
The digital model train	series	572	1 × 2764	11.75	23.50
Darkroom clock	2/90	583	1 × 27128	10.85	21.70
Video mixer	3/90	5861	1 × 2764	11.75	23.50
Four-sensor sunshine recorder	6/90	5921	1 × 27128	11.75	23.50
µP-controlled telephone exchange	10/90	5941	1 × 27128	15.30	30.60
RDS decoder	2/91	5951	1 × 2764	15.30	30.60
MIDI programme changer	4/91	5961	1 × 2764	15.30	30.60
Logic analyser	(series)		see under DISKETTES below		
Logic analyser (IBM interface)	6/91	5971	1 × PAL 16L8	8.25	16.50
MIDI-to-CV interface	2/91	5981	1 × 2764	15.30	30.60
Multifunction I/O card for PCs	7-8/91	5991	1 × PAL 16L8	8.25	16.50
Amiga mouse/joystick switch	12/91	6001	1 × GAL 16V8	8.25	16.50
Stepper motor board	6/91	6011	1 × PAL 16L8	8.25	16.50
8751 emulator (incl. system disk 5.25 in.)	3/92	6051	1 × 27C64	29.40	58.80
EMON51 (incl. course disk 1661)	(series)	6061	1 × 27256	20.00	40.00
Connect 4	12/91	6081	1 × 2764	15.30	30.60
EMON51 (incl. course disk 1681)	(series)	6091	1 × 27256	20.00	40.00
FM tuner	7/92	6101	1 × 27C256	20.00	40.00
Multi-purpose Z80 card: GAL set	6/92	6111	2 × GAL 16V8	11.15	22.30
Multi-purpose Z80 card: BIOS	6/92	6121	1 × 27128	15.30	30.60
TV test pattern generator (80C32 SBC)	3/93	6151	1 × 27256	15.30	30.60
1.2 GHz multifunction frequency meter	12/92	6141	1 × 27C256	11.45	22.90
Digital audio/visual system	12/92	6171	1 × 27C256	10.30	20.60
Digital audio/visual system (software package)	2/93	6181	EPROM, GALs and disk	30.50	61.00
PAL test pattern generator	1/93	6211	1 × GAL 20V8	9.40	18.80
Watt-hour meter	2/93	6241	1 × 27256	10.00	20.00
8751 programmer	11/90	7061	1 × 8751	46.40	92.80

DISKETTES

For pre-1990 project diskettes see the December 1992 Product Overview or contact our Dorchester office

Article/Project	Issue	Order code	Disk size	Price (£)	(US$)
Digital model train	(series)	109	5.25-inch	6.75	13.50
FAX interface for IBM PCs	6/90	119	5.25-inch (2 ×)	8.25	16.50
EPROM emulator II	9/91	129	5.25-inch	6.75	13.50
RS-232 splitter	4/90	1411	5.25-inch	6.75	13.50
Centronics ADC/DAC	5/90	1421	5.25-inch	6.75	13.50
Transistor characteristic plotting (Atari ST) (for monochrome systems only)	5/90	1431	3.5-inch	7.65	15.30
ROM-copy for BASIC computer	9/90	1441	5.25-inch	7.65	15.30
Multifunction measurement card (MMC) for PCs	2/91	1461	5.25-inch	7.65	15.30
8751 programmer	11/90	1471	5.25-inch	7.65	15.30
PT100 thermometer	11/90	1481	5.25-inch	7.65	15.30
Logic analyser: IBM software & GAL IC	7-8/91	1491	5.25-inch	19.40	38.80
Logic analyser: Atari software & GAL IC	7-8/91	1501	3.5-inch	19.40	38.80
Plotter driver (D. Sijtsma)	9/91	1541	5.25-inch	11.15	22.30
PC-controlled weather station (3)	1/92	1641	5.25-inch	7.65	15.30
8-bit I/O interface for Atari ST	4/91	1571	3.5-inch	7.65	15.30
Tektronix/Intel file converter	4/91	1581	5.25-inch	7.65	15.30
B/W video digitizer for Archimedes	7-8/91	1591	3.5-inch	11.15	22.30
Timecode interface for slide controller	9/91	1611	5.25-inch	7.65	15.30
Real-time clock for Atari ST	6/91	1621	3.5-inch	7.65	15.30
24-bit colour extension for video digitizer	11/91	1631	3.5-inch	11.15	22.30
8051/8032 assembler course (IBM version)	(series)	1661	5.25-inch	7.65	15.30
A-D/D-A and I/O for I²C bus	3/92	1671	5.25-inch	7.65	15.30
8051/8032 assembler course (Atari version)	series	1681	3.5-inch	7.65	15.30
AD232 converter	4/92	1691	5.25-inch	7.65	15.30
GAL programmer	5/92	1701	5.25-inch (3 ×)	11.15	22.30
Multi-purpose Z80 card	6/92	1711	5.25-inch	7.65	15.30
Pascal routines for MMC for PCs	10/92	1751	5.25 inch	9.70	19.40
Speech/sound memory	12/92	1771	5.25 inch	7.65	15.30
I2C opto/rolay card	2/93	1821	5.25 inch	7.65	15.30
Infra-red receiver and DTMF decoder for 80C32 single-board computer	3 & 4/93	1791	5.35-inch	7.50	15.00

SELF-ADHESIVE FRONT PANEL FOILS

Article/Project Price	Issue	Order code		(£)	(US$)
Analogue multimeter	5/89	890035-F		9.30	18.60
All-solid state preamplifier	1/90	890170-F1		19.70	39.40
		890170-F2		10.85	21.70
LF/HF signal tracer	12/89	890183-F		10.00	20.00
Video mixer	3/90	87304-F		19.40	38.80
Q meter	4/90	900031-F	•	13.50	27.00
Budget sweep/function generator	5/90	900040-F		11.75	23.50
High-current h_{FE} tester	9/90	900078-F		16.45	32.90
400-watt laboratory PSU	11/90	900082-F		20.55	41.10
Mill ohmmeter	12/90	910004-F		16.45	32.90
The complete preamp	3/91	890169-F		8.80	17.60
Wattmeter	4/91	910011-F		9.70	19.40
Universal NiCd battery charger	6/91	900134-F		6.45	12.90
Logic analyser	(series)	900094-F		10.30	20.60
Digital phase meter	6/91	910045-F		11.75	23.50
Variable AC power supply	6/91	900104-F		16.45	32.90
Timecode interface for slide controller	9/91	910055-F		8.80	17.60
Digital function generator	10/91	910077-F		10.60	21.20
4-Megabyte printer buffer	6/92	910110-F		11.45	22.90
Economy power supply	12/91	910111-F		10.60	21.20
Measurement amplifier	2/92	910144-F		8.80	17.60
CD player	1/92	910146-F		12.05	24.10
FM tuner	(series)	920005-F		13.20	26.40
LC meter	3/92	920012-F		11.45	22.90
Digital audio/visual system	(series)	920022-F1		10.00	20.00
		920022-F2		19.40	38.80
		920022-F3		28.80	57.60
Guitar tuner	6/92	920033-F		8.80	17.60
NICAM decoder	5/92	920035-F		8.25	16.50
12VDC-to-240VAC inverter	7/92	920038-F		16.15	32.30
Audio DAC	10/92	920063-F		10.00	20.00
1.2GHz multifunction frequency meter	12/92	920095-F		13.80	27.60
U2400B NiCd battery charger	2/93	920098-F		8.75	17.50

READERS SERVICES

Except in the USA and Canada, all orders, except for subscriptions and past issues (for which see below), must be sent BY POST to our Dorchester office using the appropriate form opposite. Please note that we can not deal with PERSONAL CALLERS, as no stock is carried at the editorial and administrative office.

Readers in the USA and Canada should send orders, except for subscriptions (for which see below), to Old Colony Sound Lab, Peterborough, whose full address is given on the order form opposite. All US$ prices are postpaid to customers in the fifty states except for books. Please add $2.00 for the first book and 75¢ for each additional book ordered. Canadians please add US$4.50 for the first book, and 75¢ for each additional one. Canadians may expect Canadian duty charges on shipments of any items except books.

All other customers must add postage and packing charges for orders up to £25.00 as follows: UK and Eire £1.75; surface mail outside UK £2.25; Europe (airmail) £2.75; outside Europe (airmail) £3.50. For orders over £25.00, but not exceeding £100.00, these p&p charges should be doubled. For orders over £100.00 in value, p&p charges will be advised.

SUBSCRIPTIONS & PAST ISSUES

Subscriptions and past issues, if available, should be ordered from **Worldwide Subscription Service Ltd, Unit 4, Gibbs Reed Farm, Pashley Road, TICEHURST TN5 7HE, England**. For subscriptions, use order form opposite. Prices of past issues, including postage for single copies are £2.50 (UK and Eire); £2.80 (surface mail outside UK); £3.00 (air mail Europe); and £3.75 (airmail outside Europe).

PAST ARTICLES

Photocopies of articles from January 1978 onwards can be provided, postage paid, at £1.75 (UK and Eire), £1.90 (surface mail outside UK), £2.25 (airmail Europe), or £2.50 (airmail outside Europe). In case an article is split into instalments, these prices are applicable **per instalment**. Photocopies may be ordered from our editorial and administrative offices.

COMPONENTS

Components for projects appearing in *Elektor Electronics* are usually available from appropriate advertisers in this magazine. If difficulties in the supply of components are envisaged, a source will normally be advised in the article. It should be noted that the source(s) given is (are) not exclusive — other suppliers may also be able to help.

BOOKS

The following books are currently available: these may be ordered from certain bookshops, Old Colony Sound Lab (USA and Canadian readers only), or direct from our Dorchester office.

301 Circuits	£8.95	$12.50
302 Circuits	£8.95	$12.50
303 Circuits	£9.95	$15.90
304 Circuits	£10.95	$19.95
Microprocessor Data Book	£9.95	$17.90
Data Sheet Book 2	£8.95	$16.50
Data Book 3: Peripheral Chips	£9.95	$17.95
Data Book 4: Peripheral Chips	£9.95	$17.95
Data Book 5: Application Notes	£9.95	$17.95

SHELF BOX

Elektor Electronics shelf box	£2.95	$6.00

FRONT PANELS

PROJECT	No.	Price (£)	Price (US$)
Timecode interface	910055-F	8.80	17.60
Digital function generator	910077-F	10.60	21.20
4-Megabyte printer buffer	910110-F	11.45	22.90
Economy PSU	910111-F	10.60	21.20
CD Player	910146-F	12.05	24.10
Measurement amplifier	910144-F	8.80	17.60
FM tuner	920005-F	13.20	26.40
LC meter	920012-F	11.45	22.90
Guitar tuner	920033-F	8.80	17.60
NICAM decoder	920035-F	8.25	16.50
12VDC to 240VAC inverter	920038-F	16.15	32.30
Audio DAC	920063-F	10.00	20.00
Dig. audio/visual system	920022-F1	10.00	20.00
	920022-F2	19.40	38.80
	920022-F3	28.80	57.60
1.2 GHz multifunction frequency meter	920095-F	13.80	27.60
U2400B NiCd battery charger	920098-F	8.75	17.50

EPROMS / PALS / MICROCONTROLLERS

PROJECT	No.	Price (£)	Price (US$)
Multifunction measurement card for PCs (1 × PAL16L8)	561	10.30	20.60
Video mixer (1 × 2764)	5861	11.75	23.50
Four-sensor sunshine recorder (1 × 27128)	5921	11.75	23.50
µP-controlled telephone exchange (1 × 27128)	5941	15.30	30.60
RDS decoder (1 × 2764)	5951	15.30	30.60
MIDI programme changer (1 × 2764)	5961	15.30	30.60
Logic analyser (IBM interface) (1 × PAL 16L8)	5971	8.25	16.50
MIDI-to-CV interface	5981	15.30	30.60
Multifunction I/O for PCs (1 × PAL 16L8)	5991	8.25	16.50
Amiga mouse/joystick switch (1 × GAL 16V8)	6001	8.25	16.50
Stepper motor board - 1 (1 × PAL 16L8)	6011	8.25	16.50
4-Megabyte printer buffer (1 × 2764)	6041	15.30	30.60
8751 emulator incl. system disk (MSDOS)	6051	29.40	58.80
Connect 4 (1 × 27C64)	6081	15.30	30.60
EMON51 (8051 assembler course) (1 × 27256 +disk 1661)	6061	20.00	40.00
EMON51 (8051 assembler course) (1 × 27256 +disk 1681)	6091	20.00	40.00
Multi-purpose Z80 card: FM tuner (1 × 27C256)	6101	20.00	40.00
GAL set (2 × GAL 16V8)	6111	11.15	22.30
Multi-purpose Z80 card: BIOS (1 × EPROM 27128)	6121	15.30	30.60
1.2 GHz multifunction frequency meter(1 × 27C256)	6141	11.45	22.90
Digital audio/visual system (1 × 27C256)	6171	10.30	20.60
TV test pattern generator (1 × 27256)	**6151**	**13.00**	**26.00**
DiAV system. Package: 1 × 27512; 2 × GAL; 1 × floppy disk (MSDOS)	6181	30.50	61.00
PAL test pattern generator (1 × GAL 20V8-25)	6211	9.40	18.60
Watt-hour meter (1 × 27256)	6241	10.00	20.00
8751 programmer (1 × 8751)	7061	46.40	92.80

DISKETTES

PROJECT	No.	Price (£)	Price (US$)
Multifunction measurement card (MMC) for PCs	1461	7.65	15.30
8751 programmer	1471	7.65	15.30
PT100 thermometer	1481	7.65	15.30
Logic analyser: IBM software on disk, incl. GAL	1491	19.40	38.80
Logic analyser: Atari software on disk (3.5"), incl. GAL	1501	19.40	38.80
Plotter driver (D. Sijtsma)	1541	11.15	22.30
I/O interface for Atari	1571	7.65	15.30
Tek/Intel file converter	1581	7.65	15.30
B/W video digitizer	1591	11.15	22.30
Timecode interface	1611	7.65	15.30
RTC for Atari ST	1621	7.65	15.30
24-bit colour extension for video digitizer	1631	11.15	22.30
PC controlled weather station - 3 (supersedes disks 1551 and 1561)	1641	7.65	15.30
8051/8032 Assembler course (IBM version)	1661	7.65	15.30
8051/8032 Assembler course (Atari version) (3.5")	1681	7.65	15.30
AD232 converter	1691	7.65	15.30
GAL programmer (3 disks)	1701	11.15	22.30
Multi-purpose Z80 card	1711	7.65	15.30
EPROM emulator II	129	6.75	13.50
Pascal library for MMC	1751	9.70	19.40
Speech/sound memory	1771	7.65	15.30
IR receiver and DTMF decoder for 80C32 SBC	**1791**	**7.50**	**15.00**
I2C opto/relay card	1821	7.65	15.30

PRINTED CIRCUIT BOARDS

Printed circuit boards whose number is followed by a ✣ sign are only available in combination with the associated software item, and can **not** be supplied separately. The indicated price includes the software.

OCTOBER 1992			
Audio DAC - 3	920063-3	26.45	52.90
8051 SBC	contact Suncoast Technologies		
Mains sequencer	920013	17.35	34.70
Wideband active antenna	924101	3.25	6.50
RDS demodulator	880209 ●	5.30	10.60
● Limited supply			
NOVEMBER 1992			
Printer sharing unit	920011	14.70	29.40
Sound sampler for Amiga	920074	6.75	13.50
Difference thermometer	920078	5.30	10.60
Low-power TTL-to-RS232 interface	920127	3.55	7.10
DECEMBER 1992			
Digital audio/visual system (incl. EPROM 6171)	920022✣	34.10	68.20
1.2 GHz multifunction frequency meter (incl. EPROM 6141)	920095✣	29.40	58.80
Output amplifier for ribbon loudspeakers	920135-1	19.40	38.80
	920135-2	7.95	15.90
Peak-delta NiCd charger	920147	4.10	8.20
Small projects:			
Diskette side chooser	924045	Not available	
4-digit counter	924006	Not available	
Thermocouple-to DMM interface	924052	Not available	
40-W output amplifier	924054	Not available	
IDC-to-box header adaptor	924049	6.45	12.90
Mini keyboard for Z80	924047	12.35	24.70
80C552 µP system	924071	20.00	40.00
Speech/sound memory	924012	Not available	
60-W music amplifier	924083	Not available	
Charging temperature monitor	924066	Not available	
Temperature-frequency converter	924020	Not available	
Mains power-on delay	924055	6.45	12.90
JANUARY 1993			
PAL test pattern generator (incl. GAL 6211)	920129✣	15.30	30.60
Dual video amplifier/splitter	920153	Not available	
Cross-over point detector	920165	Not available	
Multi-core cable tester			
- matrix board	926079	17.05	34.10
- slave unit	926084	6.20	12.40
- master unit	926085	8.25	16.50
FEBRUARY 1993			
U2400B NiCd battery charger	920098	8.75	17.50
Digital audio enhancer	920169	14.25	28.50
I²C opto/relay card	930004	11.00	22.00
Watt-hour meter (PCBs -1 and -2, and EPROM 6241)	920148✣	37.25	74.50
MARCH 1993			
Linear sound pressure meter	930006	7.00	14.00
Electrically isolated RS232 interface	920138	10.25	20.50
80C32 DTMF decoder	920070	Not available	

A list of all PCBs, software products and front panels available through the Readers Services is published in the March, June, September and December issues of *Elektor Electronics*.